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Proceedings of the IRE

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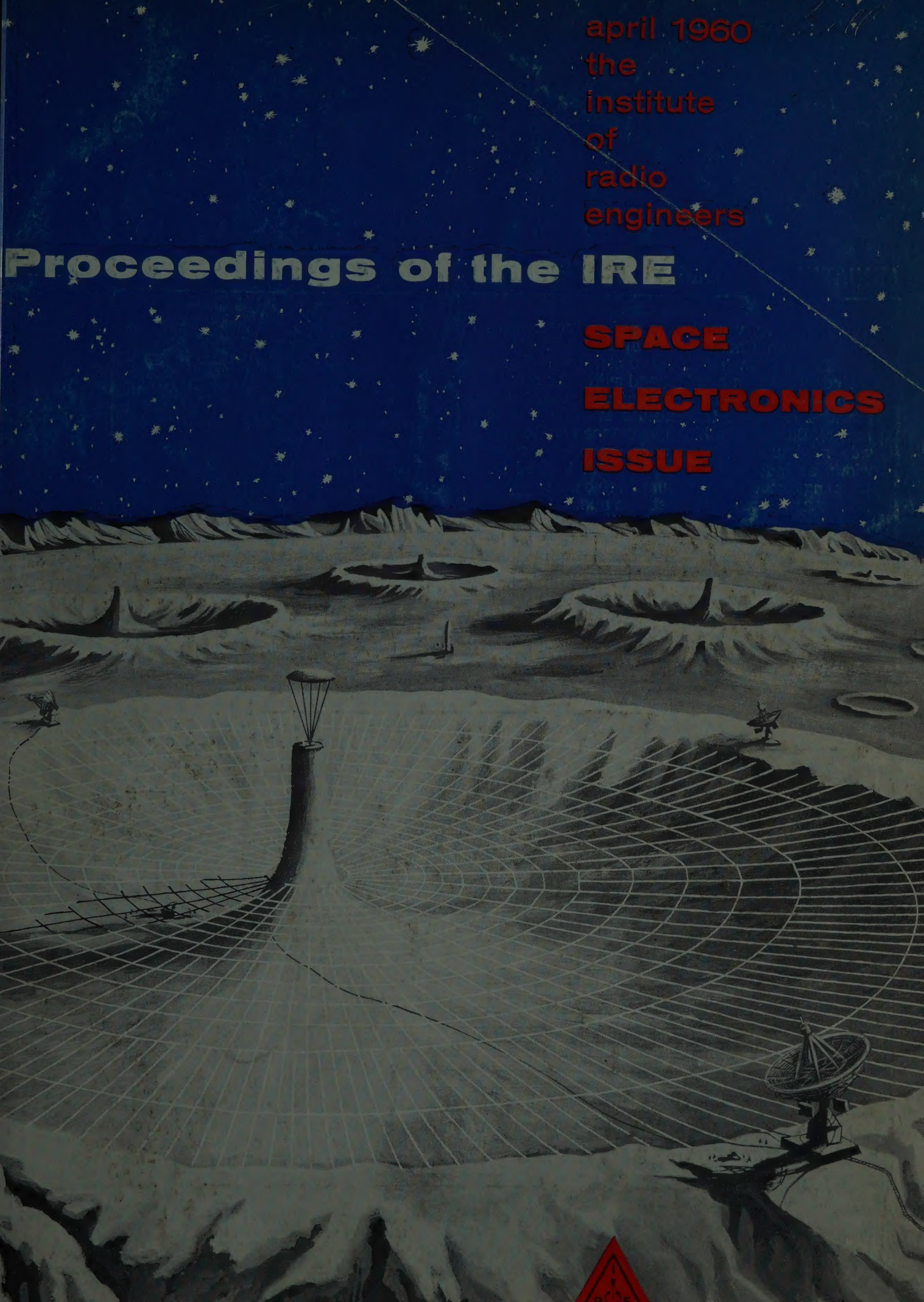

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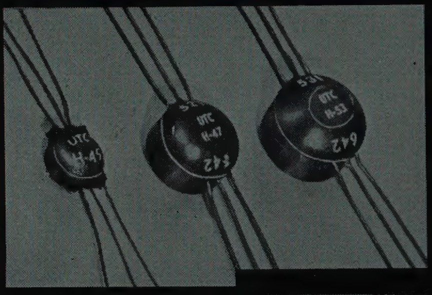
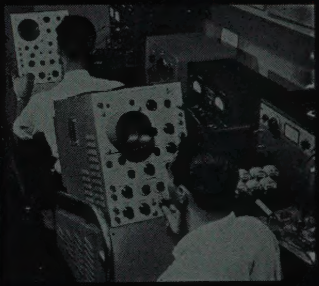


DEFINITIONS

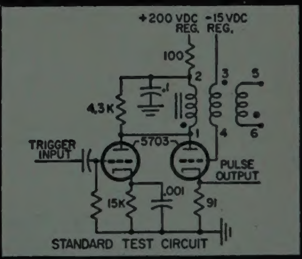
Amplitude: Intersection of leading pulse edge with smooth curve approximating top of pulse.
Pulse width: Microseconds between 50% amplitude points on leading and trailing pulse edges.
Rise Time: Microseconds required to increase from 10% to 90% amplitude.
Overshoot: Percentage by which first excursion of pulse exceeds 100% amplitude.
Droop: Percentage reduction from 100% amplitude a specified time after 100% amplitude point.
Backswing: Negative swing after trailing edge as percentage of 100% amplitude.



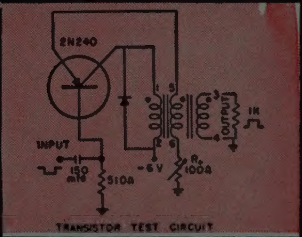
	APPROX. DCR, OHMS			BLOCKING OSCILLATOR PULSE					COUPLING CIRCUIT CHARACTERISTICS						
Type No.	1-2	3-4	5-6	Width µ Sec.	Rise Time	% Over Shoot	Droop %	% Back Swing	P Width µ Sec.	Volts Out	Rise Time	% Over Shoot	Droop %	% Back Swing	Imp. in, out, ohms
H-45	3	3.5	4	.05	.022	0	20	10	.05	17	.01	20	0	35	250
H-46	5.5	6.5	7	.10	.024	0	25	10	.10	19	.01	30	10	50	250
H-47	3.7	4.0	4	.20	.026	0	25	8	.20	18	.01	30	15	65	500
H-48	5.5	5.8	6	.50	.03	0	20	5	.50	20	.01	30	20	65	500
H-49	8	8.5	9	1	.04	0	20	10	1	24	.02	15	15	65	500
H-50	20	21	22	2	.05	0	20	10	2	27	.05	10	15	35	500
H-51	28	31	33	3	.10	1	20	8	3	26	.07	10	10	35	500
H-52	36	41	44	5	.13	1	25	8	5	23	.15	10	10	45	1000
H-53	37	44	49	7	.28	0	25	8	7	24	.20	10	10	50	1000
H-54	50	58	67	10	.30	0	20	8	10	24	.25	10	10	50	1000
H-55	78	96	112	16	.75	0	20	10	16	23	.40	5	15	20	1000
H-56	93	116	138	20	1.25	0	25	10	20	23	.6	5	10	10	1000
H-57	104	135	165	25	2.0	0	30	10	25	24	1.5	5	10	10	1000
H-60	.124	.14	.05	.05	.016	0	0	30	.05	9.3	.012	0	0	20	50
H-61	.41	.48	.19	.1	.016	0	0	30	.1	8.2	.021	0	0	15	50
H-62	.78	.94	.33	.2	.022	0	0	18	.2	7.4	.034	0	5	12	100
H-63	1.86	2.26	.70	.5	.027	2	10	20	.5	7.5	.045	0	20	25	100
H-64	3.73	4.4	1.33	1	.033	0	12	25	1	7	.078	0	15	23	100
H-65	6.2	7.3	2.22	2	.066	0	15	25	2	6.6	.14	0	10	20	100
H-66	10.2	12	3.6	3	.087	0	18	30	3	6.8	.17	0	10	20	100
H-67	14.5	17.5	5.14	5	.097	0	23	28	5	7.9	.2	0	18	28	200
H-68	42.3	52.1	14.8	10	.14	0	15	28	10	6.5	.4	0	15	30	200

Note: 0 = Negligible

Vacuum Tube Type Ratio 1:1:1



Transistor Type Ratio 4:4:1



H-45, 46, 60 thru 68 are 3/8 cube, 1 gram H-47 thru 52, 9/16 cube 4 grams H-53 thru 57, 5/8 cube 6 grams

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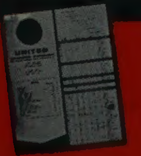

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COVER

The craters of the moon resemble the splash of a gigantic drop of water, with a high rim around the edge and a tall cone protruding from the center. This conformation has led the organizer of this special issue, Conrad Hoeppner of Radiation, Inc., to suggest that when man reaches the moon he utilize a moon crater as the framework for a giant parabolic antenna, one to two miles in diameter, to handle all communication with the earth. To produce a parabola rather than a catenary, the wires suspended from the crater rim must be uniformly weighted in the horizontal plane. This is achieved by properly selecting the size and spacing of the cross wires. Since the earth's position will appear to oscillate several degrees during each orbital cycle of the moon, provision would be made to shift the antenna beam by mounting the antenna feed structure, located atop the center cone of the crater, on tracks so that it could be moved from side to side of the antenna focal point.

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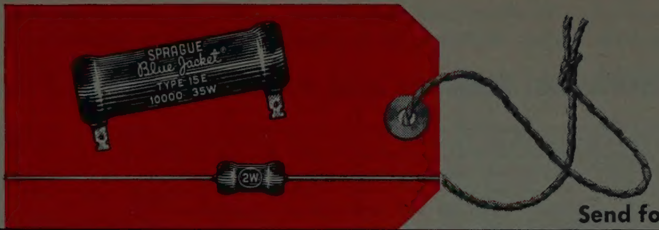
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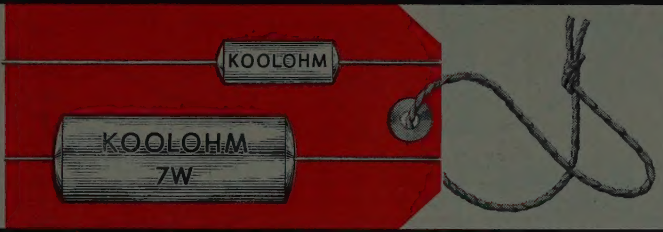
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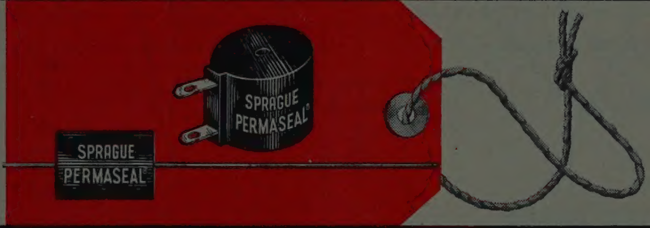
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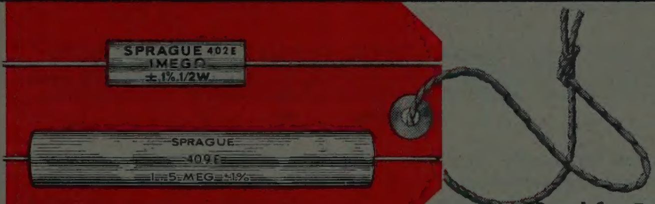
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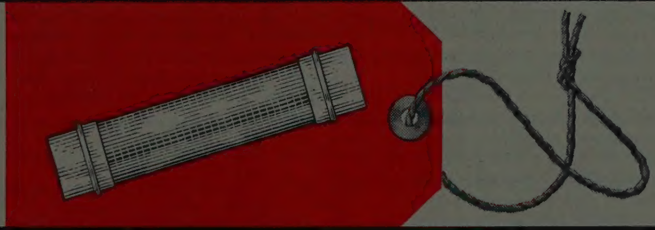

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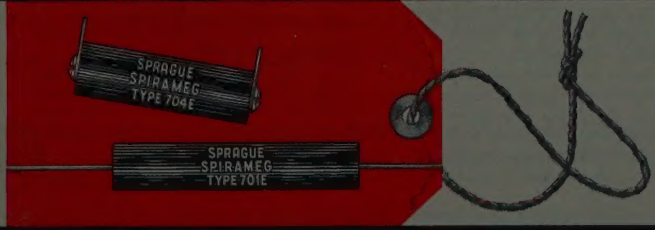
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In two articles that appeared on this page Jack Greene of our Department of Applied Electronics, discussed noise factor vs. noise temperature and antenna noise temperature. He now demonstrates the extreme importance of environment in overall receiving system sensitivity. Environment has been ignored too often when in reality, with modern low noise systems, it is often the limiting factor.

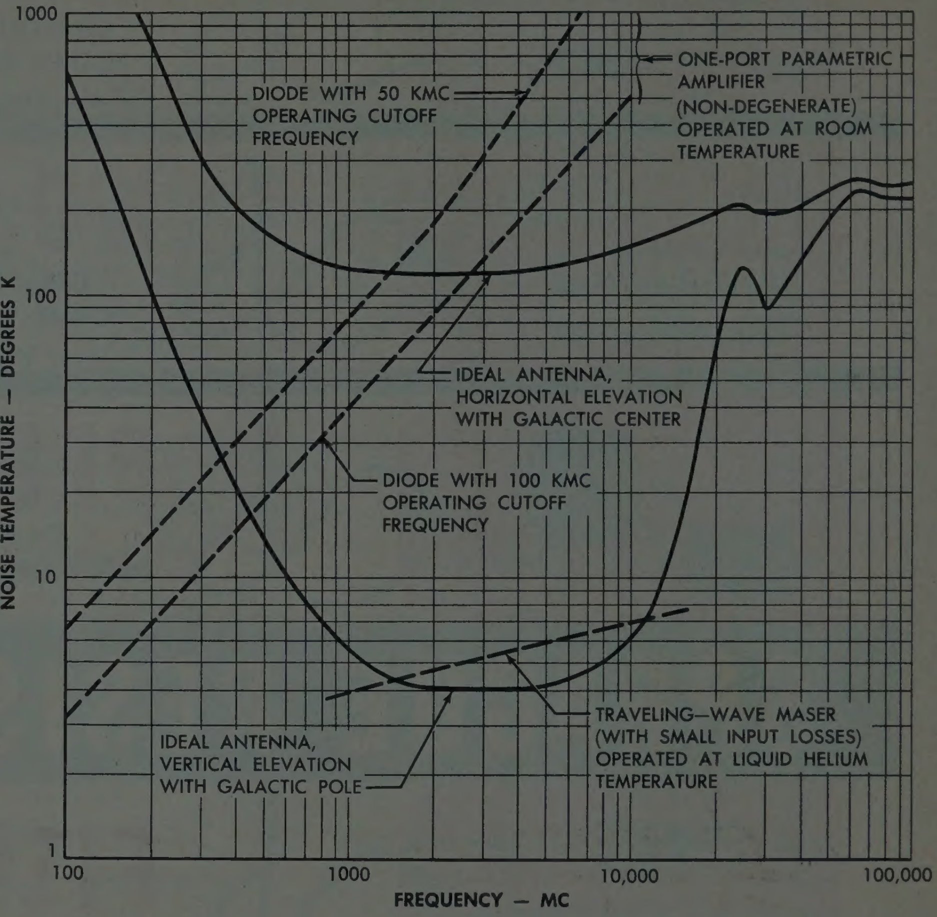
Sensitivity Limitations Set by Ideal Antennas, Masers and Parametric Amplifiers

The low noise performance of the solid-state maser and the parametric amplifier has been well established. In turn, this exceptional performance has caused a reappraisal of receiving antenna design since, in many instances, the antenna noise may determine the overall system sensitivity. Thus, it is of interest to examine the limits on antenna noise temperature present even with ideal antennas, and to further examine the limits on receiver noise temperature set by masers and parametric amplifiers.

The concepts of antenna noise temperature and effective receiver noise temperature were discussed previously.[1,2] To summarize, the sum of the antenna noise temperature and the effective receiver noise temperature determines the effective system noise temperature, and therefore the sensitivity of the complete system to received signals. Antenna noise temperature is a measure of the noise power available from the antenna terminals per unit bandwidth; it is caused by galactic noise and the absorption in the antenna environment (with its concomitant radiation) acting on the antenna gain pattern. Effective receiver noise temperature, T_e, is a term now widely used as a measure of receiver sensitivity; it is related to receiver noise factor, F, by the expression $T_e = (F - 1)290$, and represents the internally generated excess noise temperature of the receiver.

The noise temperature of an ideal antenna is readily determined if the antenna pointing angle, operating frequency, and weather conditions are specified. (A nonideal antenna with losses, spillover, and minor lobes will have reduced gain and a generally higher noise temperature; an exact knowledge of the nonideal gain pattern is required to accurately determine the noise temperature in this case.)[3] The limiting values for the noise temperature of an ideal antenna, assumed located on the earth's surface, occur as follows: (1) the minimum occurs when the antenna is pointed vertically upward (minimum atmospheric attenuation) at the galactic pole (minimum galactic noise); (2) the maximum occurs when the antenna is pointed along the horizon (maximum atmospheric attenuation) at the galactic center (maximum galactic noise). These limits are plotted in the accompanying figure as a function of frequency, under the assumption of moderate relative humidity in the atmosphere. It should be recognized that even for a fixed antenna pointing angle, the antenna noise temperature will vary with time because of the varying galactic noise component introduced by the earth's rotation, and also because of varying weather conditions.

An ideal receiver has an effective noise temperature of 0°K. Present solid-state masers approach ideal performance primarily because they are operated in liquid helium baths. (As improved maser materials become available, the stringent refrigeration requirement can possibly be relaxed without a degradation in noise performance.) The effective noise temperature obtainable from a traveling-wave maser with small input cable loss is also indicated on the accompanying figure, and in general is somewhat larger than the minimum antenna noise temperature.



The junction-diode parametric amplifier has received considerable attention because it promises low effective noise temperatures without the need for refrigeration. In fact, for a lossless diode, the noise performance of an unrefrigerated parametric amplifier would closely match that of present masers. Unfortunately, available junction diodes do have moderate losses, and therefore, their room temperature performance departs significantly from ideal performance. The noise performance expected from one-port (non-degenerate) parametric amplifiers operated at room temperature with linearly graded silicon varactors is also shown on the accompanying figure.[4] As indicated, the use of unrefrigerated parametric amplifiers would presently be restricted to the lower microwave frequencies in optimum systems. (The noise performance of the parametric amplifier using lossy diodes can be improved considerably by refrigeration;[5] however, this requires diodes specifically designed for low temperature operation, and such diodes are not yet commercially available.)

In conclusion, limits have been established for the noise temperature contributions of ideal antennas and available low noise receivers. The traveling-wave maser is a nearly optimum counterpart for an antenna with ideal characteristics, and as diode manufacturing techniques improve, unrefrigerated parametric amplifiers will seriously challenge the maser in many system applications. Finally, although the most sensitive receiving system possible is now nearly a reality with earth-bound antennas, further improvements in receiver noise temperature may become necessary for antennas located in space, where the antenna noise temperature generally decreases significantly.

1. "Noise Factor and Noise Temperature," Proc. IRE, vol. 46, p. 2A, January 1958.
2. "Antenna Noise Temperature," Proc. IRE, vol. 46, p. 4A, May 1958.
3. R. Gardner, "Final Engineering Report on Antenna Noise Temperature Study," AIL Report No. 3304-11, November 1957. (Performed under Contract DA-49-170-sc-1547.)
4. J. C. Greene and E. W. Sard, "Optimum Noise and Gain-Bandwidth Performance for a Practical One-Port Parametric Amplifier," submitted for publication in Proc. IRE.
5. M. Uenohara and W. M. Sharpless, "An Extremely Low-Noise 6-Kmc Parametric Amplifier Using Gallium-Arsenide Point-Contact Diodes," Proc. IRE, vol. 47, p. 2114-2115, December 1959.

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are the smallest of Sprague's family of film capacitors. Type 148P and 149P Yellow-Jackets are designed for compact radio receivers, test equipment, communications equipment, and similar applications. They are especially suited for transistorized and low-voltage tube circuits, as well as all other applicable circuits in which size, weight, and cost are important considerations.

Yellow-Jacket capacitor sections are of extended foil design...wound from ultra-thin, especially selected polyester film and thin gage foil under carefully controlled atmospheric conditions. They are protected against moisture by an outer wrap of polyester film. End seals are of a plastic resin which bonds securely with the film wrap in order to assure long service life.

This construction results in a light-weight capacitor of minimum size, having a distinct space advantage over metal-encased, molded, or wax-coated cardboard-case tubulars of comparable ratings.

Yellow-Jacket Type 148P (cylindrical) and 149P (semi-oval) capacitors are recommended for use in applications requiring reliable operation within the temperature range of −55 C to +85 C at rated working voltages of 100, 200, 400, and 600 volts d-c.

For complete technical data on these Yellow-Jackets, write for Bulletin 2063A to Technical Literature Section, Sprague Electric Company, 235 Marshall St., North Adams, Mass.

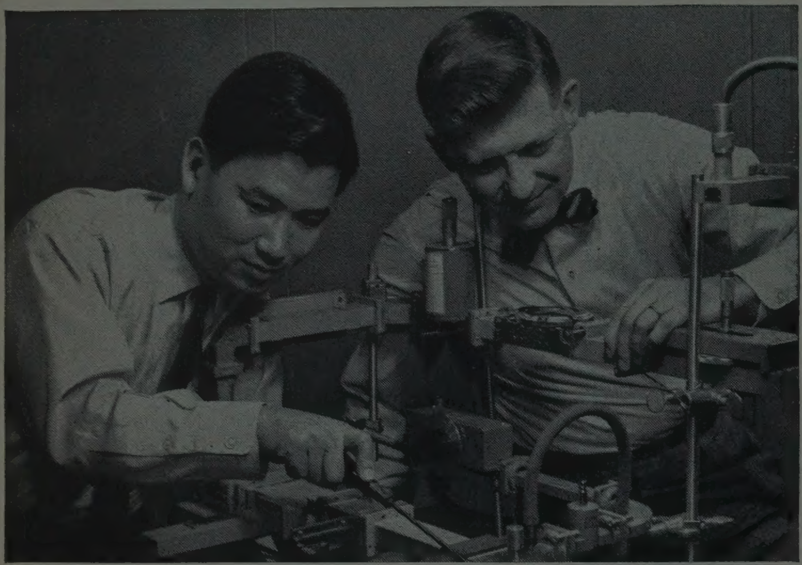
SPRAGUE®

THE MARK OF RELIABILITY

SPRAGUE COMPONENTS:

CAPACITORS • RESISTORS • MAGNETIC COMPONENTS • TRANSISTORS • INTERFERENCE FILTERS • PULSE NETWORKS
HIGH TEMPERATURE MAGNET WIRE • CERAMIC-BASE PRINTED NETWORKS • PACKAGED COMPONENT ASSEMBLIES

THE IDEA THAT GREW FOR 100 YEARS



At Bell Laboratories, M. Uenohara (left) adjusts his reactance amplifier, assisted by A. E. Bakanowski, who helped develop first suitable diode. Extremely low "noise" is achieved when certain diodes are cooled in liquid nitrogen.

First practical diode for amplifier, shown here held by tweezers, was jointly developed by A. E. Bakanowski and A. Uhlir.

How basic scientific ideas develop in the light of expanding knowledge is strikingly illustrated by the development of Bell Laboratories' new "parametric" or "reactance" amplifier.

Over 100 years ago, scientists experimenting with vibrating strings observed that vibrations could be amplified by giving them a push at strategic moments, using properly synchronized tuning forks. This is done in much the same way a child on a swing "pumps" in new energy by shifting his center of gravity in step with his motion.

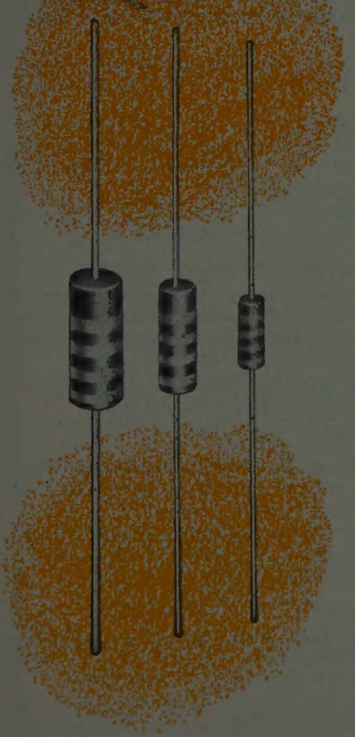
At the turn of the century, scientists theorized that *electrical* vibrations, too, could be amplified by synchronously varying the *reactance* of an inductor or capacitor. Later amplifiers were made to work on this principle but none at microwave frequencies.

Then came the middle 50's. Bell Telephone Laboratories scientists, by applying their new transistor technology, developed semiconductor diodes of greatly improved capabilities. They determined theoretically *how* the electrical capacitance of these new diodes could be utilized to amplify at *microwave* frequencies. They created a new microwave amplifier with far less "noise" than conventional amplifiers.

The new reactance amplifier has a busy future in the battle with "noise." At present, it is being developed for applications in tropospheric transmission and radar. But it has many other possible applications, as well. It can be used, for instance, in the reception of signals reflected from satellites. It is still another example of the continuing efforts to improve your Bell System communications.

BELL TELEPHONE LABORATORIES

WORLD CENTER OF COMMUNICATIONS RESEARCH AND DEVELOPMENT



how many Resistors have you soldered recently?

It's no trick today to obtain resistors that give everything you need in the way of conventional characteristics such as load life, resistance-temperature, temperature cycling, and so on.

But what a whale of a difference when it comes to "solderability"! Try the different makes for yourself and see. Whether you solder by hand or by automatic dipping, you'll find that Stackpole Coldite 70+ resistors solder lots better, lots faster and lots more surely.

Just hit 'em with solder and they stay soldered—because they're the only resistors whose leads get an extra final solder dip *in addition* to the usual tin-lead coating. You get faster production, fewer rejected assemblies. And there's less chance of trouble developing after your products reach the field.

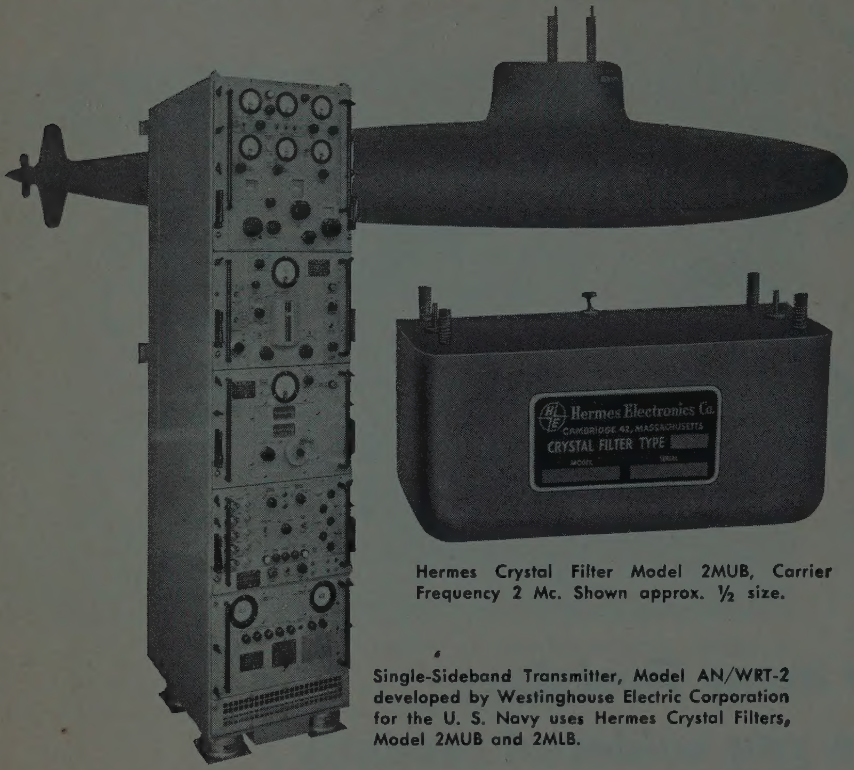
COMPARE THESE "SPECS"! — Write for Stackpole Resistor Bulletin giving complete scorecard for Coldite 70+ (cold-molded) resistors in relation to MIL as well as commercial specifications. And remember that they give you unmatched solderability in the bargain—at no extra cost!

Electronic Components Division **STACKPOLE CARBON CO.,** *St. Marys, Pa.*

Ceramag® ferromagnetic cores • Slide and Snap switches • Variable composition resistors Ceramagnet® ceramic magnets • Fixed composition capacitors • Electrical contacts Brushes for all rotating electrical equipment Hundred of related carbon, graphite, and metal powder products.

STACKPOLE
Coldite 70+®
fixed composition resistors

FIRST Navy Militarized SSB Transmitter Generates Cleaner Signal Using HERMES CRYSTAL FILTERS



Hermes Crystal Filter Model 2MUB, Carrier Frequency 2 Mc. Shown approx. ½ size.

Single-Sideband Transmitter, Model AN/WRT-2 developed by Westinghouse Electric Corporation for the U. S. Navy uses Hermes Crystal Filters, Model 2MUB and 2MLB.

Recently installed on the atomic submarine SKIPJACK (SSN585), the Westinghouse Electric AN/WRT-2 SSB Transmitter is now standard Navy equipment.

Single sideband signals are generated in the AN/WRT-2 by the selective filter method employing Hermes 2MUB and 2MLB Crystal Filters. These 2.0 Mc Crystal Filters not only offer all the basic advantages of the filter SSB generation method, but reduce the number of heterodyning stages required to translate the modulated signal to the required output frequency. The attendant decrease in unwanted signal generation results in a cleaner signal. The AN/WRT-2 is also a more reliable transmitter because fewer components are used.

In addition to the 2.0 Mc Crystal Filters, Hermes has also supplied SSB units at 87 Kc, 100 Kc, 137 Kc, 1.4 Mc, 1.75 Mc, 3.2 Mc, 6 Mc, 8 Mc, 10 Mc and 16 Mc. These Crystal Filters are presently installed in airborne HF, mobile VHF and point to point UHF SSB systems.

Whether your selectivity problems are in transmission or reception, AM or FM, mobile or fixed equipment, you can call on Hermes engineering specialists to assist in the design of circuitry and the selection of filter characteristics best suited to your needs. Write for Crystal Filter Short Form Catalog.

A limited number of opportunities are available to experienced circuit designers. Send résumé to Dr. D. I. Kosowsky.

Hermes
ELECTRONICS CO.
75 CAMBRIDGE PARKWAY, CAMBRIDGE 42, MASSACHUSETTS

Meetings with Exhibits

● As a service both to Members and the industry, we will endeavor to record in this column each month those meetings of IRE, its sections and professional groups which include exhibits.

Δ

April 20-22, 1960

SWIRECO, Southwestern IRE Regional Conference & Electronics Show, Shamrock-Hilton Hotel, Houston, Texas.

Exhibits: Mr. A. D. Seixas, SWIRECO, P.O. Box 22331, Houston, Texas.

May 2-4 1960

National Aeronautical Electronics Conference, Dayton Biltmore Hotel, Dayton, Ohio.

Exhibits: Mr. Edward M. Lisowski, General Precision Lab., Inc., Suite 452, 333 West First St., Dayton 2, Ohio.

May 2-6, 1960

Western Joint Computer Conference, Fairmont Hotel, San Francisco, Calif.

Exhibits: Mr. H. K. Farrar, Pacific Tel. & Tel. Co., 140 New Montgomery St., San Francisco 5, Calif.

May 23-25, 1960

1960 National Telemetering Conference, Miramar Hotel, Santa Monica, Calif.

Exhibits: Mr. William Van Dyke, Douglas Aircraft Co., Inc., El Segundo, Calif.

May 24-26, 1960

Seventh Regional Technical Conference & Trade Show, Olympic Hotel, Seattle, Wash.

Exhibits: Mr. Rush Drake, 1806 Bush Place, Seattle 44, Wash.

May 24-26, 1960

Armed Forces Communications & Electronics Association Convention and Exhibit, Sheraton-Park Hotel, Washington, D.C.

Exhibits: Mr. William C. Copp, 72 West 45th St., New York 36, N.Y.

June 20-21, 1960

Chicago Spring Conference on Broadcast and Television Receivers, Graemere Hotel, Chicago, Ill.

Exhibits: Mr. Stanley Hopper, Zenith Radio Corp., 6001 W. Dickens Ave., Chicago 39, Ill.

June 27-29, 1960

National Convention on Military Electronics, Sheraton-Park Hotel, Washington, D.C.

Exhibits: Mr. L. David Whitelock, BuShips, Electronics Div., Dept. of Navy, Washington, D.C.

August 1-3, 1960

Fourth Global Communications Symposium, Hotel Statler, Washington, D.C.

Exhibits: Mr. Robert O. Brady, Office of the Chief Signal Officer, U. S. Army Signal Corps, Washington, D.C.

(Continued on page 10A)

For High Accuracy—Broad Band Noise Figure Measurement

Therma-Node
KAY ELECTRIC CO.
PINE BROOK, N.J.
METER SENSITIVITY

KAY *Therma-Node*

Basic Noise Source
CAT. NO. 770

Available Noise
Accurate to ±0.1

3 Noise Heads
Cover 1 kc to 1000 mc

Portable
Can Be Operated from
117 V, 60 cycles or 24 V dc

SPECIFICATIONS

STANDARD NOISE HEAD: (Head A furnished with **Therma-Node**) covers 2 to 1000 mc; output impedance 50 ohms, unbalanced N type connectors.

Max. VSWR, variable tuned, 1.1 from 10 to 1000 mc.

Max. VSWR, fixed tuned, 1.1 from 10 to 100 mc; 1.2 from 6 to 300 mc; 1.4 from 4 to 400 mc; 2 from 2 to 500 mc.

INTERCHANGEABLE LOW FREQUENCY NOISE HEAD: (Head B) covers 1 kc to 350 mc; output impedance 50 ohms, unbalanced.

Max. VSWR 1.1 from 3 kc to 100 mc; 1.2 from 2 kc to 250 mc; 1.4 from 1 kc to 350 mc.

AMBIENT SOURCE PROBE for use with A and B Noise Heads:

Frequency range—0 to 1000 mc.

Output impedance—50 ohms, unbalanced.

Max. VSWR = 1.1

Accuracy of indicated temperature—2 EM—±1%

SELECTABLE-IMPEDANCE NOISE HEAD: (Head C) covers 0.25 to 400 mc, balanced or unbalanced output. Selectable output impedances of 50, 100 and 200 ohms are provided.

Max. VSWR—1.1 from 1 to 75 mc; 1.2 from 0.5 to 100 mc; 1.4 from 0.25 to 400 mc.

Max. VSWR difference between ambient and hot sources is 0.05 (hot and ambient sources contained in same probe).

Weight: 8 pounds in carrying case.

Dimensions: 11.5 x 8 x 4.75 inches.

Operates on 117 V, 60 cps, or 24 V dc.

Price: $495.00, f.o.b. factory.
Low Frenquency Noise Head (B): $175.00
Selected Impedance Head (C): $125.00

Through refinement of a basic noise generation technique—thermal noise from a heated resistive element—the new Kay *Therma-Node* achieves high accuracy over an extreme wide range of frequencies. *Therma-Node's* resistive element, contained in the noise head, is heated to a normal 2200°K, generating adequate noise-power for accurate noise figure measurements to 10 db. Nominal fixed available noise temperature ranging between 2000 and 2400°K may be read directly on the panel meter to 2% accuracy. A single tuning element, contained in the noise head, provides a fixed range of 2 to 500 mc, and may be tuned to extend the range to 1000 mc. An optional, interchangeable head extends measurement down to 1 kc. Ambient termination is supplied. Both heads have output impedances of 50 ohms, unbalanced. Both heads can be used without connecting coaxial cables, thus eliminating cable errors.

A selectable-impedance noise head, covering the range .25 to 400 mc, and furnishing output impedances of 50, 100 and 200 ohms, balanced and unbalanced, is available as an accessory.

The inexpensive resistive element in the *Therma-Node* noise head has a life expectancy of 10,000 hours in either intermittent or continuous service. Because the few active components in the *Therma-Node* are solid state devices, its inherent stability results in long term accuracy and freedom from maintenance.

OTHER KAY NOISE GENERATORS

Instrument & Cat. No.	Frequency Range (mc)	Noise Figure Range (db)	Output Impedance (ohms)	Price f.o.b. factory
Mega-Node® 240-B	5-220	0-16 at 50 ohms 0-23.8 at 300 ohms	unbal.—50, 75, 150, 300, ∞ bal.—100, 150, 300, 600, ∞	$365.00
Mega-Node 175-A	50-500	0-19	balanced—300	$365.00
Mega-Node 403-A	3-500	0-19	unbalanced—50	$365.00
Mega-Node 3000	1-3000	0-20	unbalanced—50	$790.00
Rada-Node® 600-A	5-400	0-23.8 depending on impedance	unbalanced as specified	$1495.00
	10-3000	0-20	unbal. nom. 50	$1965.00
	1120-26,500	15.28 or 15.8	waveguide	†
*Microwave Mega-Nodes**	1120-26,500	15.28 or 15.8	waveguide	$175.00 to $595.00

† Price varies with Microwave Mega-Node used as accessory.
* Ideally suited for noise figure measurement in radar communication.

WRITE FOR NEW KAY CATALOG

KAY ELECTRIC COMPANY
Dept. I-4 Maple Avenue, Pine Brook, N.J. CApital 6-4000

TOWER FOOTING INSULATORS FOR SELF-SUPPORTING RADIATORS
MAST BASE INSULATORS
RADIO GUY INSULATOR

LAPP ANTENNA TOWER INSULATORS

We at Lapp are mighty proud of our record in the field of tower insulators. Over 30 years ago, the first insulated broadcasting tower was erected—on Lapp insulators. Since then, most of the large radio towers in the world have been insulated and supported by Lapp insulators. Single base insulator units for structures of this type have been design-tested to over 3,500,000 pounds.

A thorough knowledge of the properties of porcelain, of insulator mechanics and electrical qualities has been responsible for Lapp's success in becoming such an important source of radio insulators. Write for description and specification data on units for any antenna structure insulating requirement. Lapp Insulator Co., Inc., Radio Specialties Division, 223 Sumner Street, LeRoy, N. Y.

Lapp

Meetings with Exhibits

(Continued from page 8A)

August 23-26, 1960
WESCON, Western Electronic Show and Convention, Ambassador Hotel & Memorial Sports Arena, Los Angeles, Calif.
Exhibits: Mr. Don Larson, WESCON, 1435 LaCienega Blvd., Los Angeles, Calif.

September 19-21, 1960
National Symposium on Space Electronics & Telemetry, Shoreham Hotel, Washington, D.C.
Exhibits: Mr. Leon King, Jansky and Bailey, 1339 Wisconsin Ave., N.W., Washington, D.C.

October 3-5, 1960
Sixth National Communications Symposium, Hotel Utica & Utica Municipal Auditorium, Utica, N.Y.
Exhibits: Mr. R. E. Bischoff, 19 Westminster Road, Utica, N.Y.

October 10-12, 1960
National Electronics Conference, Hotel Sherman, Chicago, Ill.
Exhibits: National Electronics Conference, Inc., 228 North La Salle St., Chicago 1, Ill.

October 24-26, 1960
East Coast Aeronautical & Navigational Electronics Conference, Lord Baltimore Hotel & 7th Regiment Armory, Baltimore, Md.
Exhibits: Mr. R. L. Pigeon, Westinghouse Electric Corp., Air Arm Div., P.O. Box 746, Baltimore, Md.

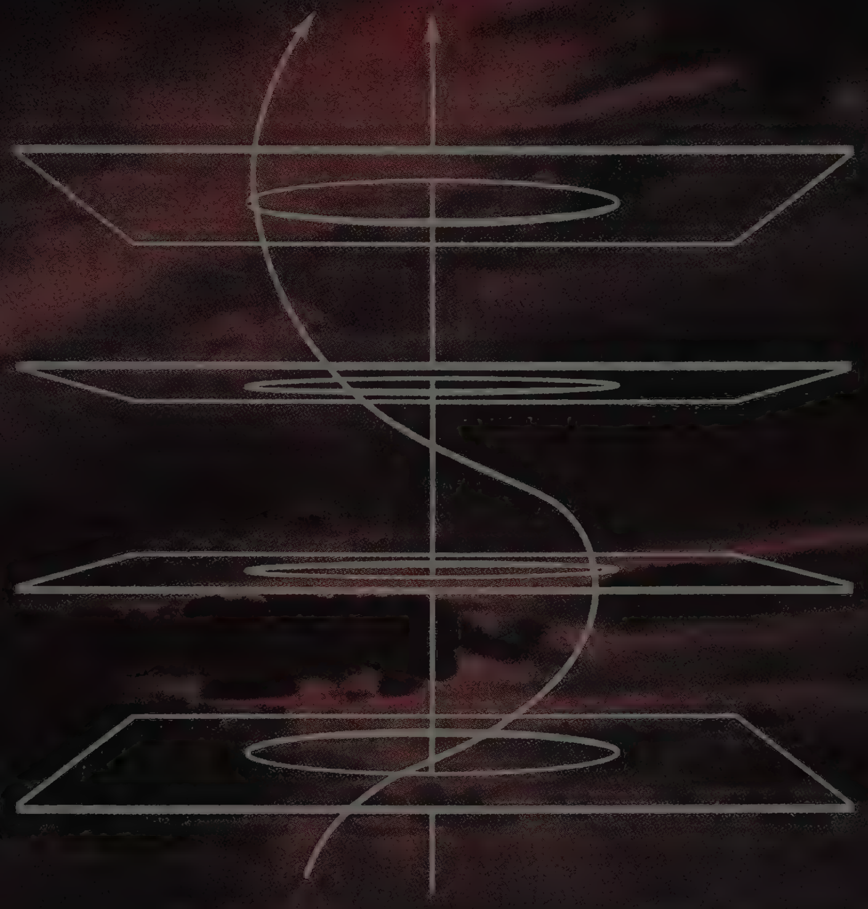
Oct. 31-Nov. 2, 1960
13th Annual Conference on Electrical Techniques in Medicine & Biology, Sheraton-Park Hotel, Washington, D.C.
Exhibits: Mr. Lewis Winner, 152 West 42nd St., New York 36, N.Y.

November 14-16, 1960
Mid-America Electronics Convention (MAECON), Municipal Auditorium, Kansas City, Mo.
Exhibits: Mr. John V. Parks, Bendix Aviation Corp., P.O. Box 1159, Kansas City 41, Mo.

November 15-17, 1960
Northeast Electronics Research & Engineering Meeting (NEREM), Boston Commonwealth Armory, Boston, Mass.
Exhibits: Miss Shirley Whitcher, IRE Boston Office, 73 Tremont St., Boston, Mass.

Note on Professional Group Meetings: Some of the Professional Groups conduct meetings at which there are exhibits. Working committeemen on these groups are asked to send advance data to this column for publicity information. You may address these notices to the Advertising Department and of course listings are free to IRE Professional Groups.

THE G. C. DEWEY CORPORATION
For scientific research and development
with intellectual integrity



SCIENTIFICALLY ORIENTED *to achieve significant results for government and civilian agencies—in the broad areas of defense physics . . . weapons system concepts and evaluation . . . electronic technology—this organization applies creative scientific thinking . . . to advanced problems of nuclear and plasma physics . . . to unique computer programming techniques . . . to original electronic design criteria . . . to systems analyses that meet measure of effectiveness standards.*

THE G. C. DEWEY CORPORATION, 202 East 44th Street, N. Y. C. 17

Physicists, mathematicians and engineers with advanced degrees,
write for information on professional career opportunities.

For original equipment, direct replacement

NEC tubes with new doped-nickel cathode

6R-P10

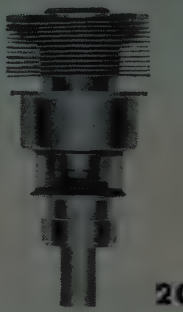
6R-R8

Both tube series described here use NEC's new doped-nickel cathode core material. This 10-year development increases emission without raising operating temperature. Oxide evaporation rate is lower than any known core material. Operating data show tube life is extended up to 50%.

WIDE-BAND AMPLIFIER TUBES: Development began seven years ago with the 6R-R8, which was used in Japan's first microwave link. A modification, 6R-R8C, with very low distortion factor, is used in coaxial amplifiers. 6R-P10 Power Amplifier Pentode, with high mutual conductance and small capacitance, is designed for larger power output.

Type	Name	Cathode Rating		Screen and Plate Supply Voltage Eb (V) Ec2 (V)	Plate Current Ib (mA)	Trans-conductance Gm (μ℧)	Capacitances		Interchangeable Tubes
		Volts Ef (V)	Amp. If (A)				Input μμF	Output μμF	
6R-R8	Sharp-Cutoff Pentode	6.3	0.3	150	13	12,500	7.8	3.2	with WE404A
6R-R8C	Sharp-Cutoff Pentode	6.3	0.3	150	13	12,500	7.3	3.2	with WE404A
6R-P10	Power Amplifier Pentode	6.3	0.5	150	36	13,500	10.5	2.7	——

2C40



2C39B

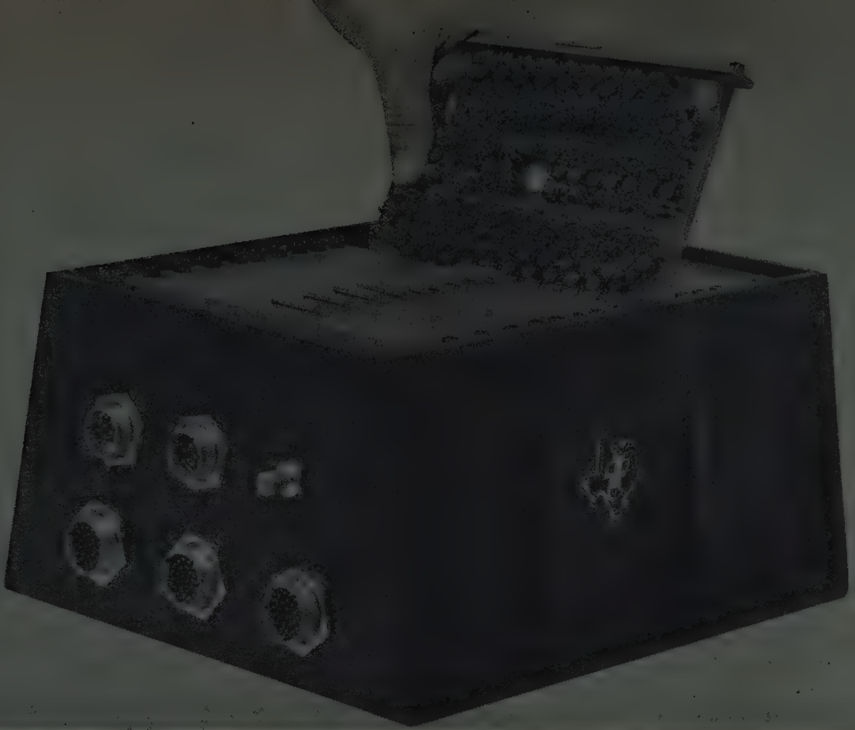
DISC-SEALED TRIODES: NEC designed the first disc-sealed tube in 1939, giving NEC many years of experience in the design and manufacture of this type of microwave tube. Each is a direct replacement under all circumstances for the corresponding type. The NEC tube will give longer life, an especially important advantage in repeater stations.

Type	Use	Cathode Rating		Maximum Plate Voltage Ev (V)	Maximum Plate Dissipation Pp (W)	Power Po (W)	Maximum Frequency f (MC)
		Voltage Ef (V)	Current If (A)				
2B22	Detector	6.3	0.75	150	—	—	1200
2C39A	Amplifier Oscillator (Continuous)	6.3	1.0	1000	100	15	2500
2C40	Amplifier Oscillator (Continuous)	6.3	0.75	500	4	0.75	3370
2C43	Amplifier Oscillator (Continuous & Pulse)	6.3	0.9	3500*	12	1.0	3370
5861	Amplifier Oscillator (Continuous)	6.3	0.4	350	10	0.5	3700

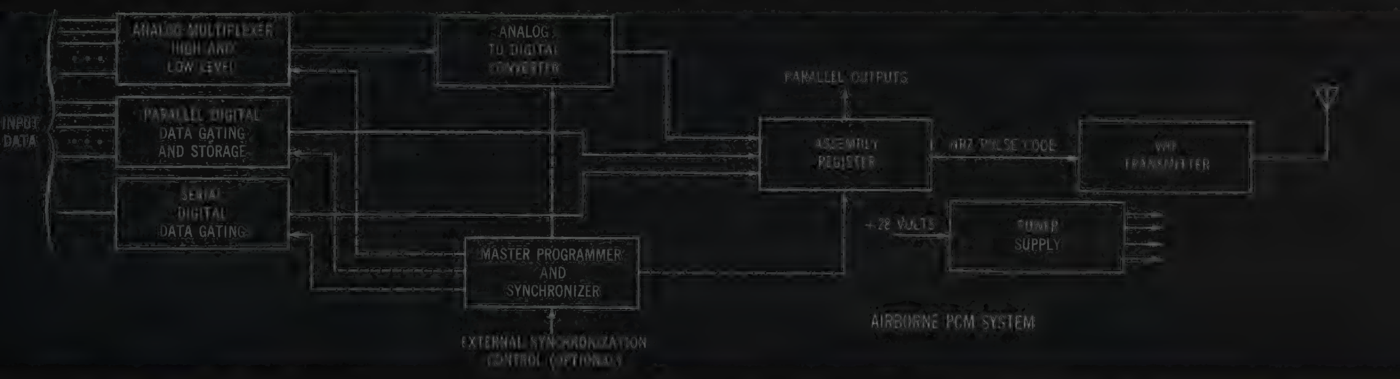
* Pulse plate Voltage (eb)

Please write for specification sheets.

NEC *Nippon Electric Company Ltd.* **Tokyo, Japan**
COMPONENTS / SYSTEMS



NEW FROM TEXAS INSTRUMENTS!


INPUT DATA
ANALOG MULTIPLEXER HIGH AND LOW LEVEL
PARALLEL DIGITAL DATA GATING AND STORAGE
SERIAL DIGITAL DATA GATING
ANALOG TO DIGITAL CONVERTER
MASTER PROGRAMMER AND SYNCHRONIZER
EXTERNAL SYNCHRONIZATION CONTROL (OPTIONAL)
PARALLEL OUTPUTS
ASSEMBLY REGISTER
NRZ PULSE CODE
VHF TRANSMITTER
+28 VOLTS
POWER SUPPLY
AIRBORNE PCM SYSTEM

High capacity, ½ cu ft, PCM telemetry system ...with system drift nulled out!

Now you can get the benefits of digital techniques — accuracy, speed and reliability — in compact pulse code modulation telemetry systems from Texas Instruments. The 25-pound, ½ cubic foot package shown uses only solid state devices, and may be used to drive any of the compact TI transmitters dictated by the application. The system multiplexes and encodes 64 analog channels; and processes five 8-bit parallel digital data channels plus a serial digital data channel at a nominal bit rate of 200 kc.

A key feature of the system is its high-speed analog multiplexer which handles low- and high-level data, or a combination of both, with *only a single low-level amplifier*. Overall accuracy of the system is ±0.25% — made possible by a unique bi-directional servo loop that nulls out system drift. The system is packaged in individual modules so that it can be rapidly modified to fit the needs of other missiles or space vehicles.

PCM is only one of several advanced telemetry projects at Texas Instruments. Others include the development and/or production of analog systems and equipments for Bomarc, Corvus, Pershing, Minuteman, Titan, Centaur, and Project Mercury.

For detailed information about PCM telemetry or other TI missile electronic system capabilities, please contact SERVICE ENGINEERING:

TEXAS INSTRUMENTS
INCORPORATED
APPARATUS DIVISION
6000 LEMMON AVENUE
DALLAS 9, TEXAS

IRE News and Radio Notes

CURRENT IRE STATISTICS

(As of February 29, 1960)

Membership—79,646
Sections*—105
Subsections*—27
Professional Groups*—28
Professional Group Chapters—261
Student Branches†—184

* See March, 1960 issue for a list.
† See October, 1959 issue for a list.

Calendar of Coming Events and Authors' Deadlines*

1960

- **6th Nuclear Congress, N. Y. Coliseum. New York, N. Y., Apr. 4–8.**
- **14th Spring Tech. Conf., Cincinnati, Ohio, Apr. 12–13.**
- **Conf. on Automatic Tech., Sheraton-Cleveland Hotel, Cleveland, Ohio, Apr. 18–19.**
- **Int'l Symp. on Active Networks and Feedback Systems, Engrg. Soc. Bldg. Auditorium, New York, N. Y., Apr. 19–21.**
- **Int'l Symp. on Active Networks and Feedback Systems, Polytechnic Inst. of Brooklyn, Brooklyn, N. Y., Apr. 19–21.**
- **1960 SWIRECO (Southwestern IRE Regional Conf. and Electronics Show), simultaneously with the Nat'l. PGME Conf., Houston, Texas, Apr. 20–22.**
- **Natl. Aeronautical Electronics Conf., Dayton-Biltmore and Miami Hotels, Dayton, Ohio, May 1–3.**
- **URSI-IRE Spring Mtg., Sheraton Park Hotel and NBS, Washington, D. C., May 2–5.**
- **Western Joint Computer Conf., San Francisco, Calif., May 2–6.**
- Symp. on Graduate Programs in Bio-Medical Engrg., Univ. of Vermont, Burlington, May 5–6.
- **PGMTT Natl. Symp., San Diego, Calif., May 9–11.**
- **Electronic Components Conf., Hotel Washington, Washington, D. C., May 10–12.**
- **Natl. Telemetering Conf., Miramar Hotel, Santa Monica, Calif., May 23–25.**
- **7th Reg. Tech. Conf. & Trade Show, Olympic Hotel, Seattle, Wash., May 24–26.**
- **6th Radar Symp., Ann Arbor, Mich., June 1–3.**
- Inst. on Recent Advances in Solid State Devices, Marquette Univ., Milwaukee, Wis., June 1–2.
- 10th Ann. Conv. of Soc. of Women Engrs., Benjamin Franklin Hotel, Seattle, Wash., June 9–11.
- **Radio Frequency Interference Symp., Washington, D.C., June 13–14.**

* DL=Deadline for submitting abstracts.

(Continued on page 15A)

ANNOUNCE SYMPOSIUM ON SUPERCONDUCTIVE TECHNIQUES FOR COMPUTING SYSTEMS

A Symposium on Superconductive Techniques for Computing Systems will be held on May 17–19, 1960, sponsored by the Information Systems Branch, Office of Naval Research. The Symposium will be held in the Department of Interior Auditorium on C Street, between 18th and 19th Streets, N.W., Washington, D. C.

The purpose of this Symposium is to bring together the scientists and engineers currently engaged in cryogenic device research in order to present a complete picture of the present status of the applications of superconductivity to computers, computing systems, and information processing devices. Invited papers emphasizing device aspects have been solicited from most of the organizations in the United States which are now involved in research of this type. Approximately 16 to 18 papers will be presented at the Symposium including those by representatives of the following activities: Burroughs Corp., Duke University, General Electric Company, International Business Machines Corp., Lincoln Laboratory of M.I.T., Arthur D. Little, Inc., University of North Carolina, Radio Corporation of America, Rutgers University, Space Technology Laboratory, and Sperry-Rand Corporation. Attendance is open to all interested technical personnel.

Individuals may receive further information and a preliminary Symposium program when available by contacting: Miss Josephine Leno, Code 430A, Office of Naval Research, Washington 25, D. C.

RUSSIAN TECHNICAL JOURNAL AVAILABLE IN TRANSLATION

The Society for Industrial and Applied Mathematics announces the appearance of the first issue of Volume IV of *Theory of Probability and Its Applications.* This is a complete translation into English of the corresponding issue of the Russian journal *Teoriya Veroyatnostei i ee Primeneniya.* During 1960 the Society will publish separately translations of all four issues of Volume IV (1959), will begin the translation of Volume V, and will publish in bound form full translations of Volumes I (1956), II (1957) and III (1958). It is expected that by early 1961 translations will be appearing within four months of publication of the Russian original.

Theory of Probability and Its Applications is a quarterly journal devoted, as its title indicates, to research papers in probability and statistics, and to related applications in physics and communication. For the most part, the journal has reported the work of Russian authors; its translation offers the first access to this material in a Western language.

The Society for Industrial and Applied Mathematics has been assisted in this translation project by a grant from the National Science Foundation.

Subscriptions to *Theory of Probability and Its Applications* are being offered at $19.00 for four current issues (one year) ($9.50 to members of the Society, add $3.90 for subscriptions outside of the U. S. and Canada). Inquiries may be addressed to the Society at Box 7541, Philadelphia 1, Pa.

AIR FORCE MARS EASTERN NET CHOOSES SPEAKERS FOR APRIL

The April schedule of the Air Force MARS Eastern Technical Net, operating Sundays from 2 to 4 P.M. EST, 3295–7540–15,715 kc, has been built around the three subjects for which hundreds of requests have been received over the past few months. Through the cooperation of the Philco Technological Center it is possible to present the following series:

April 3	"Television and Scanning Techniques in the Field of Medical Electronics," J. F. Fisher, Engineering Section Manager, Government and Industrial Div., Philco Corp.
April 10	"Applications of Tunnel Diodes," J. A. Ekiss, Project Engineer, Special Projects Group, Lansdale Tube Co., Philco Corp.
April 17	Recess Date.
April 24	"Transistor Circuit Considerations," H. W. Merrihew, Director, Special Course Preparation, Philco Technological Center.

URSI SPRING MEETING PROGRAM ANNOUNCED

The program of invited papers for the Spring Meeting of URSI Combined Technical Session has been selected. In keeping with the times, it emphasizes problems associated with space. The following papers will be given:

"The Transmission Media in Space Communication," *I. Katz, Applied Physics Lab., The Johns Hopkins University, Baltimore, Md.*

"The NASA Communications Satellite Program," *L. Jaffe, Communications Satellite Program, National Aeronautics and Space Administration, Washington, D. C.*

"Advanced Research Project Agency's Satellite Communication Systems," *E. E. Harriman, ARPA, Washington, D. C.*

"Information Techniques in Space Communication," *G. E. Mueller, Space Technology, Labs. Inc., Los Angeles, Calif.*

"The Role of Ranging Experiments in Space Exploration," *S. W. Golomb, Jet Propulsion Lab., California Institute of Technology, Pasadena, Calif.*

AIEE PLANS SPACE CONFERENCE AT DALLAS, TEXAS, APRIL 11–13

The first special conference on electrical engineering in space technology sponsored by the American Institute of Electrical Engineers will be held at the Baker Hotel, Dallas, Tex., April 11–13, General Chairman B. J. Wilson, of the Naval Research Laboratory, Washington, D. C., has announced.

"Electrical aspects of space technology have become increasingly significant in fields of aero and astro-dynamics, propulsion merging with and overlapping those of metallurgy, and atomic and solid-state physics," Mr. Wilson said. "Those working in these fields have become aware of this new significance of electrical science and engineering and see the need to utilize the existing capabilities here to solve space problems."

The conference, which is expected to attract leading engineers and scientists, will explore electrical engineering aspects of communications, feedback control, electrical energy conversion and instrumentation in space. It will stress space requirements and the capabilities of electrical science and engineering to fulfill these needs.

Specifically, the conference will deal with the national space picture, the space age viewpoint on communications, control, electrical energy and instrumentation; specific missions, requirements and problems of some United States missiles and space vehicles and payloads and developments in equipment and component capabilities for space applications.

Several of the most prominent figures in the space technology and in the political world have been invited to address the two and one-half day conference.

The IRE and the American Rocket Society will participate in the conference.

13th ANNUAL ETMB CONFERENCE ISSUES CALL FOR PAPERS

Plans are now being made for the 13th Annual Conference on Electrical Techniques in Medicine and Biology, which this year will be held in Washington, D. C., October 31, November 1 and 2, 1960, at the Sheraton-Park Hotel.

As in the past, the meeting will be sponsored by the Joint Executive Committee in Medicine and Biology representing the Institute of Radio Engineers, American Institute of Electrical Engineers and the Instrument Society of America.

The theme of the conference will be the application of electronic techniques to analytical instumentation.

Scheduled for discussion are the exploration and application of new electrical and physical techniques for quantification of specific materials, determining molecular structure or providing recognizable characterization of materials of biological interest. Typical subject areas here are:

1) Polarography and specific electrodes; perm-selective membrane methods.
2) Electrical generation of titrant coulometric methods.
3) Nuclear and electron magnetic resonance analysis procedures.
4) Dielectric dispersion and high-frequency titration.
5) Mass spectrometry.
6) Microwave spectroscopy.

Papers representing original contributions in these related fields are invited.

Abstracts (200–500 words in length, which can be accompanied by supplementary illustrations) with author's name, company affiliation and position title, business and home address, telephone contact, and brief biographical sketch, should be submitted in double-spaced typewritten form (in triplicate) on or before July 1 to the Program Chairman, George N. Webb, Room 547—CSB, Johns Hopkins Hospital, Baltimore 5, Md.

RQC SYMPOSIUM SOLICITS PAPERS

Abstracts of papers to be submitted for the Seventh National Symposium on Reliability and Quality Control, jointly sponsored by the IRE, the AIEE and the ASQC, must be received by May 16, 1960. The Symposium is scheduled for January 9–11, 1961, at the Bellevue-Stratford Hotel, Philadelphia, Pa.

Ten copies of the abstract, of not more than 800 words, are required. It should include the title of the paper (not to exceed 50 letters, including spaces), the author's name, position and affiliation. In the case of more than one author, please indicate who will present the paper. Ten copies of a biographical sketch (suitable for publication in the PROCEEDINGS) of each author must accompany the abstract.

Authors will be notified of acceptance by June 27, 1960. Final papers will be due October 10, 1960. Submit abstracts and biographical sketches to R. E. Kuehn, IBM Owego, Owego, N. Y.

NBS PREPARES BIBLIOGRAPHY ON RADIO PROPAGATION

In the past few years the Central Radio Propagation Laboratory has been engaged in preparing several bibliographies on tropospheric propagation of UHF, VHF, and SHF radio waves. It is believed that these bibliographies have presented a comprehensive listing of the pertinent literature concerning tropospheric propagation up through 1958.

Because of the favorable reception of the previous bibliographies, and to fill a long-felt need, it has been decided to extend the scope of this bibliography to include all fields of radio propagation.

Since this is an undertaking of considerable proportions, it would be greatly appreciated if any interested reader would forward a list of pertinent references in radio propagation, radio astronomy, radio properties of the ionosphere, etc., to the editor: Malcolm Rigby, Editor, Meteorological Abstracts and Bibliography, P. O. Box 1736, Washington 13, D. C.

Calendar of Coming Events and Authors' Deadlines*

(Continued from page 14A)

Chicago Spring Conf. on Broadcast and Television Receivers, Graemere Hotel, Chicago, Ill., June 20–21.

Conf. on Standards and Electronic Measurements, NBS Boulder Labs., Boulder, Colo., June 22–24.

Workshop on Solid State Electronics, Purdue Univ., Lafayette, Ind., June 23–24.

Natl. Conv. on Mil. Elec., Sheraton Park Hotel, Washington, D. C., June 27–29.

Cong. Intl. Federation of Automatic Control, Moscow, USSR, June 25–July 9.

Int'l Conf. on Electrical Engrg. Education, Sagamore Conf. Center, Syracuse Univ., Syracuse, N. Y., Jul.

7th Ann. Symp. on Computers and Data Processing, Stanley Hotel, Estes Park, Colo., July 28–29.

4th Global Communications Symp., Hotel Statler, Washington, D. C., Aug. 1–3.

WESCON, Los Angeles Mem. Sports Arena, Los Angeles, Calif., Aug. 23–26, (DL*: May 1, R. G. Leitner, WESCON Bus. Office, 1435 So. La Cugna Blvd., Los Angeles 35, Calif.)

URSI 13th Gen. Assembly, Univ. of London, London, Eng., Sept. 5–15.

Joint Automatic Control Conf., M.I.T., Cambridge, Mass., Sept. 7–9.

Space Electronics and Telemetry Conv. and Symp., Shoreham Hotel, Washington, D.C., Sept. 19–22.

Industrial Elec. Symp., Sheraton Cleveland Hotel, Cleveland, Ohio, Sept. 21–22.

Sixth Natl. Communications Symp., Hotel Utica and Utica Municipal Aud., Utica, N. Y., Oct. 3–5. (DL*: June 1, B. H. Baldridge, 25 Bolton Rd., New Hartford, N. Y.)

Natl. Elec. Conf., Hotel Sherman, Chicago, Ill., Oct. 10–12. (DL*: May 1960 Prof. T. F. Jones, Jr., School of E.E., Purdue Univ., Lafayette, Ind.)

Symp. on Space Navigation, Deshler-Hilton Hotel, Columbus, Ohio, Oct. 19–21.

East Coast Conf. on Aero & Nav. Elec., Baltimore, Md., Oct. 24–26.

5th Ann. Conf. on Nonlinear Magnetics and Magnetic Amplifiers, Bellevue-Stratford Hotel, Philadelphia, Pa.

Electron Devices Mtg., Hotel Shoreham, Washington, D. C., Oct. 27–29.

13th Ann. Conf. on Elec. Tech. in Med. and Bio., Sheraton Park Hotel, Washington, D. C., Oct. 31, Nov. 1–2.

Radio Fall Mtg., Hotel Syracuse, Syracuse, N. Y., Oct. 31, Nov. 1–2.

Mid-Amer. Elec. Conv., Hotel Muehlebach, Kansas City, Mo., Nov. 15–16.

1960 NEREM (Northeast Electronics Res. & Engrg. Mtg.), Boston, Mass., Nov. 15–17.

PGVC Ann. Mtg., Sheraton Hotel, Philadelphia, Pa., Dec. 1–2.

Eastern Joint Computer Conf., New Yorker Hotel, New York, N.Y., Dec.

* DL = Deadline for submitting abstracts.

VLF Transmissions Begun By U. S. Naval Observatory

The Navy Department announces that Naval Radio Station NBA at Balboa, Panama Canal Zone has begun transmitting precise time signals and constant frequency on VLF (Very Low Frequency), 18 kilocycles per second. At present the signals are transmitted continuously from 0800 to 1600 Eastern Standard Time Monday through Friday. The periods of transmission will be lengthened later as the need arises.

This service was inaugurated in order to meet the needs of the Navy for precise time signals and constant frequency, and to provide means for accurate timing needed in satellite tracking stations spread over the globe.

Experiments conducted by the U. S. Naval Observatory and the Royal Greenwich Observatory in England have shown that time signals transmitted in the very low frequency part of the radio frequency spectrum, such as the 16 kilocycles per second of Station GBR at Rugby, England, and the 18 kc used at the high power transmitter at station NBA in the Panama Canal Zone, can be received over long distances to an accuracy of five ten-thousandths of a second. Recent experiments conducted by the Naval Research Laboratory indicate that an accuracy of time measurement of one ten-thousandth of a second might be achieved. The Naval Observatory, Naval Research Laboratory, and Naval Communications are conducting experiments to improve the accuracy of the service.

Atomic clocks of the cesium-beam type are used to measure the frequency of the transmissions and to maintain the frequency constant to about 1 part in 10 billion. The frequency of cesium used is 9 192 631 770 cycles per second. This experimental value was determined jointly by the U. S. Naval Observatory, Washington, and the National Physical Laboratory, Teddington, England.

Additional details of the NBA broadcasts are contained in Time Service Notices Numbers 6, 7, and 8 of the U. S. Naval Observatory. Further information may be obtained by writing to the Superintendent.

Optical Society of America Translates Russian Journal

The Optical Society of America has announced the translation of *Optika I Spektroskopiya* (*Optics and Spectroscopy*) beginning with the January, 1959 issue. This leading scientific journal of the USSR Academy of Sciences publishes the work of Russian scientists in all branches of optics and spectroscopy, including x-ray, ultraviolet, visible, infrared and microwave, thin layer optics, filters, detectors, diffraction gratings, electroluminescence, thermal radiation backgrounds, infrared polarizers, and their other applications in science and industry.

Optics and Spectroscopy is available free to all members of the Optical Society of America as part of their membership privileges, and also to all subscribers to the *Journal of the Optical Society of America* who are not members of the Society. Membership dues, including both *Optics and Spectroscopy* and the *Journal of the Optical Society of America*, are $13.00 per year. Write to Dr. Mary E. Warga, Executive Secretary, Optical Society of America, 1155 Sixteenth St., N.W., Washington, D. C. for the membership application form. Nonmember subscriptions, which include both of the above journals, are $25.00 per year. Subscriptions should be sent directly to the American Institute of Physics, 335 E. 45 St., New York 17, N. Y.

Army MARS Technical Net Announces April Speakers

Continuing with its mission of disseminating technical knowledge by radio communication, the First U. S. Army MARS SSB Technical Net announces its speaker program for April. The net operates each Wednesday evening at 9 P.M. EST on 4030 kc upper sideband.

Arrangements have been made by Ted Mathieson, A4FJ, the state MARS Director of Virginia, to use First Army tapes in setting up a Second Army Technical Net with its nucleus around the Washington D. C. area.

Speakers for April include:

April 6 "Filter Design and Applications," J. L. Prather, Instructor, Radio Div., USASCS, Fort Monmouth, N. J.

April 13 "New Semi-Conductors for High Frequency Circuits," W. A. McCarthy, Chief Applications Engineer, Semi-Conductor Div., Raytheon Mfg. Co., Boston, Mass.

April 20 "Modern Trends in Electronic Instrumentation," W. A. Knoop, Jr., Partner, Gawler-Knoop Co., Roselle, N. J.

April 26 "Tacan and Similar Aircraft Navigation Systems," W. Loebel, Project Engineer, Olympic Radio and TV Division of the Siegler Corp., Long Island City, N. Y.

OBITUARIES

Philip R. Coursey (A'17)(L) died recently at the age of 67.

Born in 1892 in Wood Green, Middlesex, England, he received the B.S. degree, with honors, in engineering from University College, London, in 1913. After graduation, he was Professorial Assistant in the Electrical Engineering Department and Research Laboratory at University College, lecturing and doing research work in electrical engineering and radiotelegraphy. In 1912 he became a consulting engineer for wireless telegraphy and related fields to the Dubilier Electrical Syndicate, Ltd., London. At the time of his retirement two years ago he was Chief Engineer of Dubilier Condenser Co., London, England.

Mr. Coursey was a Fellow of the Physical Society of London, a member of the Wireless Society of London, and a member of the IEE in England. He was the author of several technical papers.

George Morton Cummings (A'32–M'46–SM'53), former chairman of the St. Louis, Mo., Section of the IRE, died recently at the age of 64.

Born March 20, 1895 in Sedalia, Mo., he attended grade school and high school there, and later attended Washington University Night School. From 1912 to 1915 he was employed by Union Electric Light and Power Co., and for a year after that he worked for the Wagner Electric Corp. In 1916 he began working for Southwestern Bell Telephone Co., in St. Louis. He was employed in the Engineering Department for many years, having been appointed Demonstrations Engineer for several states in 1951. In this capacity he supervised operation and maintenance of equipment used in exhibits and developed and designed new equipment.

Mr. Cummings served the St. Louis Section of the IRE in various capacities, as Vice-Chairman, Secretary-Treasurer, Program Chairman and Chairman of the Nominations Committee. In 1948 he was elected Chairman of the Section.

Dr. D. O. McCoy (A'46–SM'56–F'59), Chairman of the Cedar Rapids Section of the IRE, died recently at the age of 48.

D. O. McCoy

He was born March 18, 1911 in LaPorte City, Iowa. His formal academic preparation was done at the Iowa State University, where he received the B.S. degree in Mechanical Engineering in 1934, the M.S. in Physics in 1936, and the Ph.D. in Applied Physics in 1938. He contributed to the application of radio to Naval aircraft systems through his research at the Naval Research Laboratory from 1938 to 1947. He served the Laboratory as head of the Research Section of the Airborne Radio Division and later head of the Radar Section of Radio Division III. In 1947 he joined the Research Division of the Collins Radio Company, where his contributions to a number of fields were numerous and significant. Among his many activities was the direction of research and development in the fields of radar, electronic countermeasures, Doppler and phase-locked receiver techniques, radio astronomy, radio-celestial navigation, antennas, and mechanical engineering. He served on the RTCA subcommittee responsible for the study of the distance measuring system (DME) to be recommended for civil use. He was awarded the IRE Fellow Award for his guidance of the research and development leading to the radio sextant, an all-weather navigation system which utilizes the radio emission of the sun and moon.

Dr. McCoy was a member of Pi Mu Epsilon, Sigma Xi, the American Physical Society, and the Institute of Navigation.

ABOUT WATERLOO . . .

Shortly after Waterloo, the Duke of Wellington received a letter, postmarked St. Helena. It was from Napoleon. It read: "Excellency: I was amused to hear your recent remark that 'The Battle of Waterloo was won on the playing fields of Eton.' To have won an engagement in Belgium from a field in England, you must have been further back of the battle lines than I thought.

*"The real reasons for my defeat were two, and Eton was neither. In the first place, the radar broke down for two hours in the heat of battle. Not even a Napoleon can be expected to make radar work without Bomac tubes.**

"But I might easily have defeated you, faulty tubes and all, had I not been persuaded to partake of a bottle of Scotch on the evening before the battle. I have reason to suspect my drink was tainted. At any rate, on the day of Waterloo, I did not display my usual energy and decisiveness.

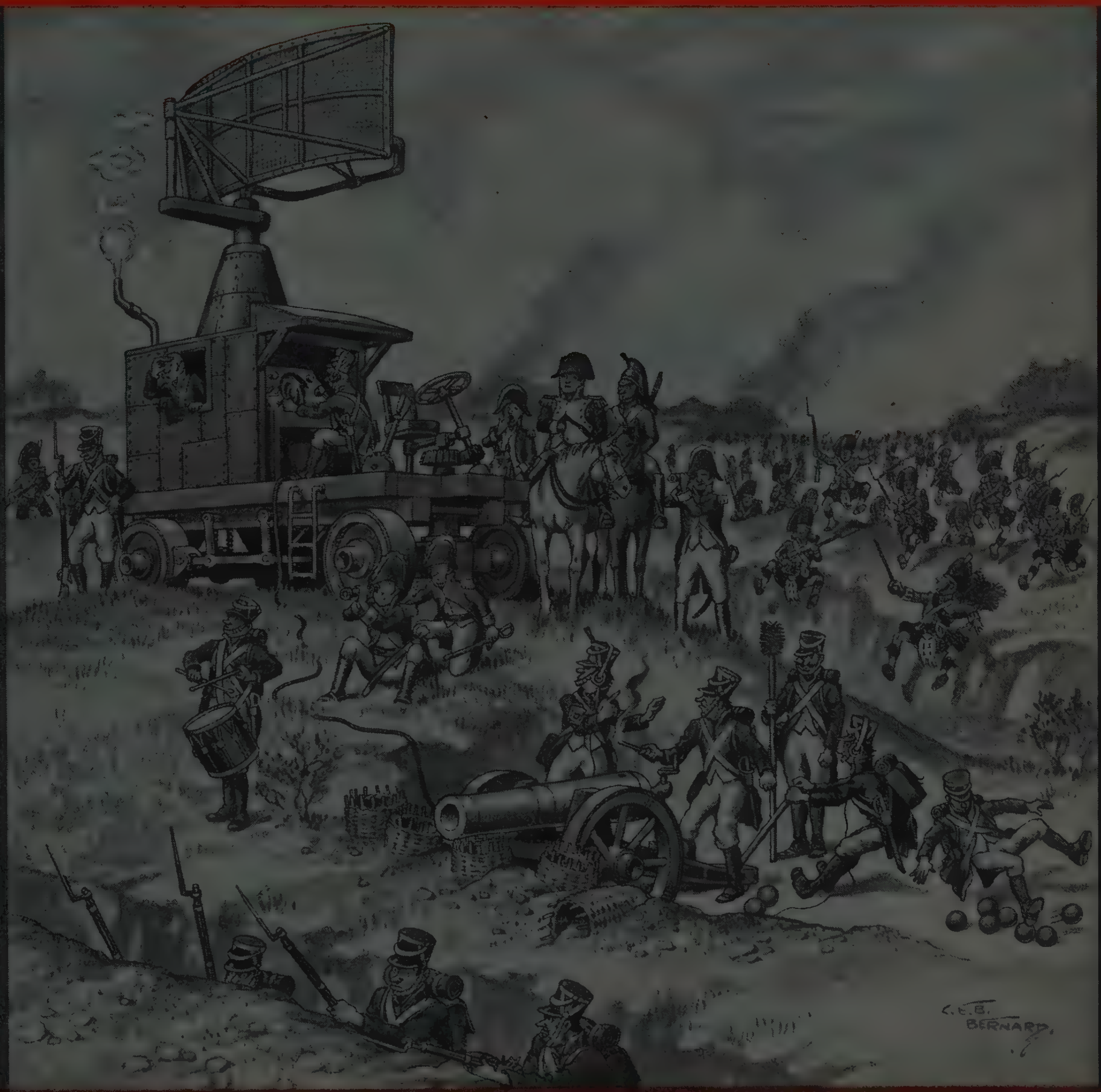
"It appears, in short, that you owe the battle to a bottle. (Signed,) N."

The Emperor received a brief reply by return boat. It read: "Excellency: In view of the fact that your loss at Waterloo appears to have been less a matter of Eton than of Drinking, I am withdrawing my original statement. I have released the following in its place, which I here submit for your approval:

'You can mix Scotch and Water
And Water and Scotch
But don't whatever you do
Make the mistake Napoleon did,
And mix Scotch and Waterloo.'

(Signed,) Wellington."

No. 19 of a series . . . **BOMAC LOOKS AT RADAR THROUGH THE AGES**



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Leaders in the design, development and manufacture of TR, ATR, Pre-TR tubes; shutters; reference cavities; crystal protectors; silicon diodes; magnetrons; klystrons; duplexers; pressurizing windows; noise source tubes; high frequency triode oscillators; surge protectors.

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1960 IRE INTERNATIONAL CONVENTION RECORD

The IRE International Convention Record, containing all available Convention papers, will be published in ten parts in July, 1960.

Professional Group members are entitled to purchase the Part sponsored by the Professional Group to which they belong at the special PG rate indicated below. Other parts may be purchased at the IRE member rate.

IRE members may purchase any part at the IRE member rate. However, if a member applies for membership in the appropriate Professional Group at the same time that he places his order, he will be entitled to the PG rate.

Nonmembers and libraries may place orders at the nonmember and library rates, respectively. Individuals who apply for IRE membership at the time they place their orders are entitled to the IRE member rate.

Subscription agencies are entitled to the library rate.

Clip out the order form and return it, with remittance, to the Institute of Radio Engineers, Inc., 1 East 79 St., New York 21, N. Y. In ordering, be sure to refer to the proper columns for subjects and prices.

Part	Sessions	Subject and Sponsoring IRE Professional Group	Prices for Members of Sponsoring Professional Group (PG), IRE Members (M), Libraries and Sub. Agencies (L), and Nonmembers (NM)			
			PG	M	L	NM
1	38, 46, 53	Antennas & Propagation	$0.70	$1.05	$2.80	$3.50
2	33, 40, 41, 48, 49	Circuit Theory Electronic Computers	1.00	1.50	4.00	5.00
3	8, 16, 23, 32, 39	Electron Devices Microwave Theory & Techniques	1.00	1.50	4.00	5.00
4	1, 9, 17, 25	Automatic Control Information Theory	0.80	1.20	3.20	4.00
5	24, 28, 29, 30, 37, 44, 50	Communications Systems Space Electronics & Telemetry	1.20	1.80	4.80	6.00
6	7, 18, 22, 27, 31, 34, 35, 42	Component Parts Industrial Electronics Production Techniques Reliability and Quality Control Ultrasonics Engineering	1.40	2.10	5.60	7.00
7	11, 12, 19, 20, 26, 36	Audio Broadcast & Television Receivers Broadcasting	1.00	1.50	4.00	5.00
8	4, 6, 15, 43, 51, 52	Aeronautical & Navigational Electronics Radio Frequency Interference Vehicular Communications Military Electronics	1.00	1.50	4.00	5.00
9	2, 10, 14, 21, 47, 54	Medical Electronics Nuclear Science Instrumentation	1.00	1.50	4.00	5.00
10	3, 5, 13, 45	Engineering Writing & Speech Engineering Management Human Factors in Electronics	0.80	1.20	3.20	4.00
		Complete Set (10 parts)	$10.00	$15.00	$40.00	$50.00

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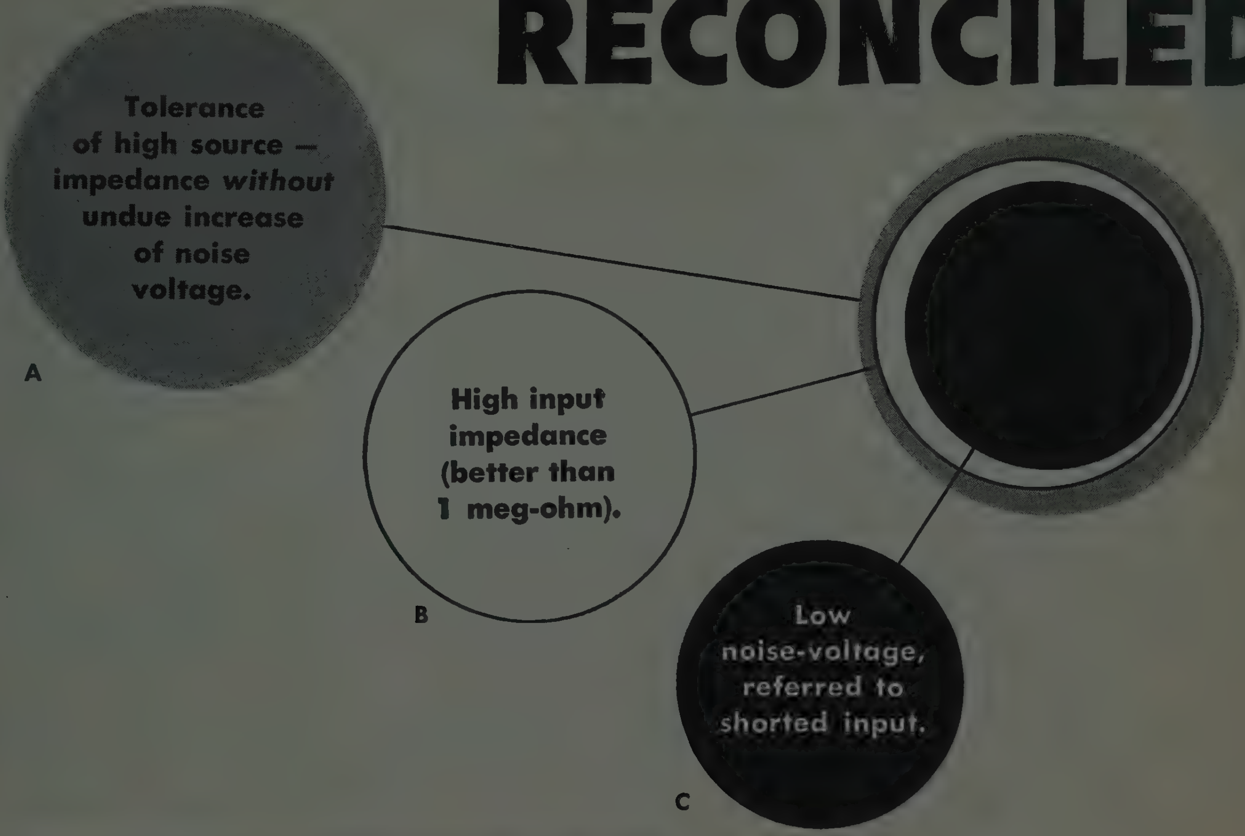

Tolerance of high source — impedance *without* undue increase of noise voltage.

A

High input impedance (better than 1 meg-ohm).

B

Low noise-voltage, referred to shorted input.

C

↓VS-66A→	Parameter	Z Source	←VS-68A ↓
0 to 40, db, 5 steps	Gain	—	10 to 80 db, 8 steps
10 meg	Z_{Input}	—	selectable 10 k or 10 meg
20 cps to 2 mc	Freq. Rg.	—	20 cps ot 180 kc
0.7 uV RMS	$e_{N\text{-}in}$, B=10 kc	0 ohms	0.2 uV, 10k input & 1.4 uV, 10 meg input
0.75 uV RMS	"	100 ohms	0.3 uV, 10k input & 1.5 uV, 10 meg input
0.8 uV RMS	"	500 ohms	0.8 uV, 10k input & 1.6 uV, 10 meg input
0.85 uV RMS	"	1 k	1.3 uV, 10k input & 1.7 uV, 10 meg input
1.9 uV RMS	"	10 k	4.2 uV, 10k input & 3.6 uV, 10 meg input
5.5 uV RMS	"	100 k	— — — 11.0 uV, 10 meg input
10 uV RMS	"	1 meg	— — — 20.0 uV, 10 meg input

The three amplifier characteristics illustrated (A, B and C) are often considered contradictory, even irreconcilable. True, there are amplifiers which exhibit low-noise voltage, referred to the shorted input terminals. Fact is, our VS-64A still holds the record in this field — less than 0.5 mu V RMS over a 60 kc pass band. Also, there are many high input impedance amplifiers. But never before have low-noise voltage characteristics and high input impedance been combined with a highly incompatible third stipulation: tolerance of high source impedance, without spoilage of premium shorted-input noise performance, by insertion of a comparatively large source impedance.

Two new Millivac low-noise amplifiers — the VS-66A and VS-68A — *combine all 3* of these most desirable characteristics to meet the demands of the Space Age for extended communications and control ranges.

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1960 Spring Technical Conference

HOTEL ALMS, CINCINNATI, OHIO, APRIL 12–13, 1960

Tuesday Morning, April 12

Opening Address: *F. C. Reith, President, Crosley Div., and Vice President of Avco Corp., Cincinnati, Ohio.*

Session 1A

Panel: Missile and Space Vehicle Re-entry.

Chairman: *Major General M. C. Demler, Assistant Deputy Chief of Staff, USAF; Director, Res. and Dev.*

"The Theory of Hypersonic Re-entry," *Dr. R. Hermann, University of Minnesota, Director of Rosemount Research Lab., Minneapolis, Minn.*

"The Engineering Approaches to the Re-entry Problem," *Dr. F. R. Ridell, Avco Corp., Research and Advanced Development Div., Wilmington, Mass.*

"Plasma Sheet Characteristics About a Hypersonic Vehicle," *Dr. H. A. Lew, General Elec. Co., Missile and Space Systems Div., Philadelphia, Pa. and V. A. Langelo, Space-Science Labs., Moorestown, N. J.*

"Performance Capabilities of Lifting Hypervelocity Vehicles," *Dr. R. B. Hildenbrand, Advanced Systems Research, Aero-Space Div., Boeing Airplane Co., Seattle, Wash.*

Session 1B

Panel: Can Computers Out-Think People?

Chairman: *J. C. Elms, Vice President and General Manager, Avco Corp., Crosley Div., Cincinnati, Ohio.*

"One Plus One is Not Necessarily Two—A Tutorial Paper on Number Systems," *A. C. Cook, Application Engineer, General Electric Co., Cincinnati, Ohio.*

"Resistor-Transistor Logic in a Large Scale Digital Computer," *J. N. Conway, General Precision Equipment Corp., Librascope Div., Burbank, Calif.*

"An All-Transistor Control Computer for Ballistic Missiles," *W. E. Lane, The Martin Co., Orlando, Fla.*

"A Macro-Module Computer," *S. Schneider, The Burroughs Corp., Paoli, Pa.*

Tuesday Afternoon—Session 2A

Panel: Why Will Tunnel Diodes Revolutionize Electronics?

Chairman: *Dr. G. Bruck, Senior Staff Scientist, Avco Corp., Crosley Div., Cincinnati, Ohio.*

"A Report on the State of the Art" *E. O. Johnson, RCA, Semiconductor and Materials Div., Somerville, N. J.*

Panel Discussion: *L. Cuccia, RCA, Harrison, N. J., N. J. Golden, Hoffman Electronics Corp., El Monte, Calif.; Dr. C. S. Kim, General Electric Co., Syracuse, N. Y.; R. Pritchard, Texas Instruments, Inc., Dallas, Texas; Dr. A. Pucel, Raytheon Co., Waltham, Mass., and Dr. E. R. Steele, Hughes Aircraft Corp., Newport Beach, Calif.*

Session 2B

Panel: The Use of Data Processing in Bio-Electronics.

Chairman: *Colonel J. P. Stapp, M.D., Director of the Aero-Space Medical Lab., WADD, USAF ARDC; Past President of the American Rocket Society.*

"Selectively Monitoring Physiological Data," *M. A. McLennon, Chief, Bio-Electronics Section, Bio-Physics Branch, Aero-Space Medical Lab., WADD.*

"Implications of Some Recent Work with Skin Resistance," *Captain V. H. Thaler, USAF (MSC), Aero-Space Medical Lab., WADD.*

"A Biological Dynamic Energy Transducer," *Dr. J. R. Mundie, M.D., Bio-Acoustics Branch, Aero-Space Medical Lab., WADD.*

"The Contributions of Biology to Electronic Design," *Major J. E. Steele, M.D., USAF, MC.*

Wednesday Morning, April 13

Session 3A

Panel: Electronic Data Processing Comes of Age (Building Blocks).

Chairman: *R. G. Counihan, Manager, Micro-Machine Development, IBM Corp., Kingston, N. Y.*

"Magnetic Tape Techniques in Data Processing," *B. Haslam, Avco Corp., Crosley Div., Electronics Research Labs., Boston, Mass.*

"High Speed Logic Using Low-Cost Mesa Transistors," *R. Lohman, RCA, Semiconductor and Materials Div., Somerville, N. J.*

"A Switching Applicaiton for PN-PN Transistors in Data Processing Machines," *P. A. Bunyar, IBM Corp., Kingston, N. Y.*

IBM Color Film: "The X-15-Man into Space."

"Circuit Techniques Used in the LARC Computer," *L. M. Spandorfer, Sperry Rand Corp., Remington Rand Univac Div., Philadelphia, Pa.*

Session 3B

Panel: Objectives and Accomplishments of the International Geophysical Year.

Chairman: *Dr. J. Kaplan, University of California, Los Angeles; Chairman of the U. S. National Committee of the 1957–58 IGY, National Academy of Sciences.*

"The Use of Rockets and Satellites as Tools to Aero-Space Research During the IGY," *Dr. R. W. Porter, General Electric Co., Corporate Headquarters, New York, N. Y.*

"Towards Our Knowledge of the Earth," *Dr. G. P. Woollard, Department of Geology, University of Wisconsin, Madison, Wis.*

"Physical Phenomena of the Cis-Lunar Space," *Dr. F. Johnson, Lockheed Aircraft Corp., Palo Alto, Calif.*

"Oceanography and the IGY," *Dr. R. R. Revelle, Scripps Institute of Oceanography..*

Luncheon Speaker: *Major General M. C. Demler; Assistant Deputy Chief of Staff, USAF; Director, Res. and Dev.*

Wednesday Afternoon—Session 4A

Panel: Inertial Guidance for Missile and Space Vehicle Applications.

Chairman: *Dr. C. S. Draper, Director, Instrumentation Lab., M.I.T., Cambridge, Mass., Chairman, U. S. Inventor's Council.*

"Problems of Precision Inertial Velocity Measurements," *Dr. J. H. Slater, Autonetics Div., American Aviation, Inc., Downey, Calif.*

"Inertially Instrumented Interplanetary Guidance," *H. Winter, Bell Aircraft Corp., Avionics Div., Inertial Development Labs., Niagara Falls, N. Y.*

"The Application of Inertial Guidance Techniques to Stabilization in Orbit," *J. J. Klien, Lockheed Aircraft Corp., Missile Systems Div., Sunnyvale, Calif.*

"Inertial Guidance—Present and Future," *R. C. Berendsen, General Electric Co., Defense Electronics Div., Pittsfield, Mass.*

"Inertially Derived Rate of Descent Measuring System," *R. K. Smyth, J. W. Montooth, and D. D. Farnum, Autonetics Div., North American Aviation, Inc., Downey, Calif.*

Session 4B

Panel: Electronic Data Processing Comes of Age (Systems).

Chairman: *T. H. Bonn, Senior Staff Scientist, Sperry Rand Corp., Remington Rand Univac Div., Philadelphia, Pa.*

"The 117L Program," *M. Ehlers, Beckman Instrument Corp., Chicago, Ill.*

"Electronic Data Processing at the Atlantic Missile Range with Emphasis on the Azusa Tracking System," *R. J. Pepple, RCA Data Reduction Center, AF Missile Test Center, Cape Canaveral, Fla.*

"A Radar Simulator Design Using Binary Digital Techniques," *Dr. R. J. McNair, Avco Corp., Crosley Div., Cincinnati, Ohio.*

"COED—A Device for the Simulation of Man/Machine Operations," *R. H. Johnson, Bendix Aviation Corp., Systems Div., Ann Arbor, Mich.*

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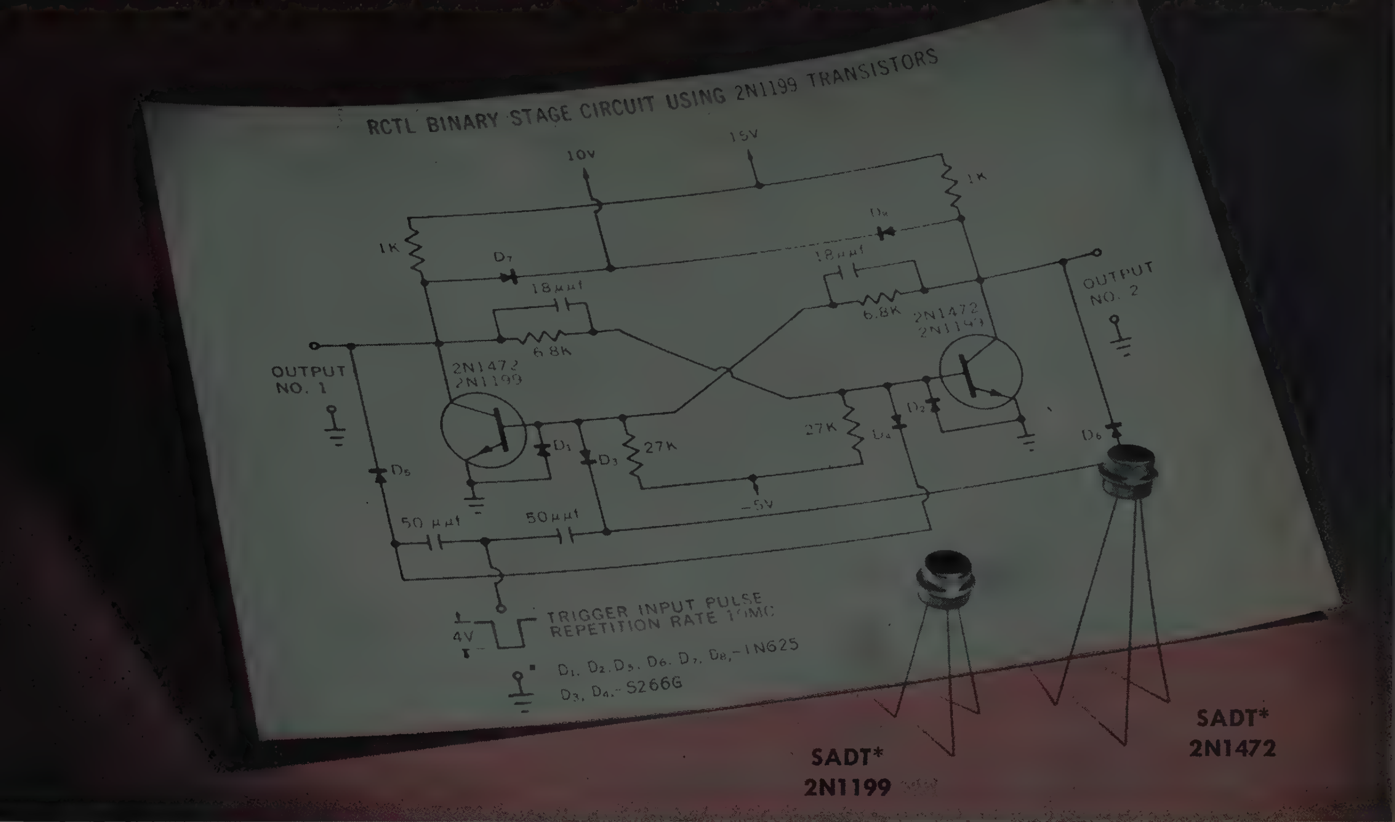
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15V
1K
1K
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6.8K
OUTPUT NO. 1
OUTPUT NO. 2
2N1472
2N1199
2N1472
2N1199
D₁
D₃
D₂
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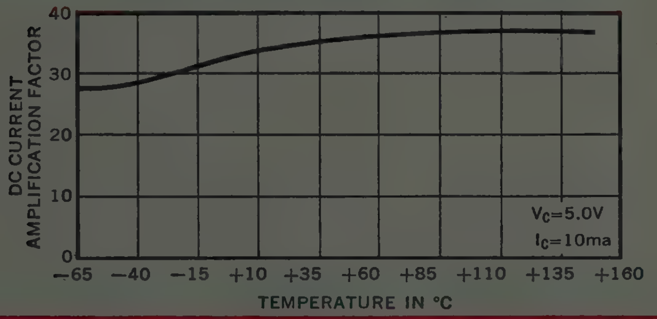
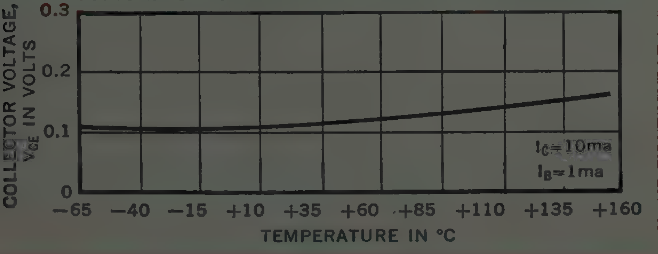
TRIGGER INPUT PULSE
REPETITION RATE 10MC
D₁, D₂, D₅, D₆, D₇, D₈,−1N625
D₃, D₄,−S266G
SADT*
2N1199
SADT*
2N1472

PHILCO HIGH FREQUENCY NPN SILICON TRANSISTORS OFFER EXCEPTIONALLY LOW SATURATION VOLTAGE

TYPICAL CURRENT GAIN VS. TEMPERATURE



TYPICAL SATURATION VOLTAGE VS. TEMPERATURE

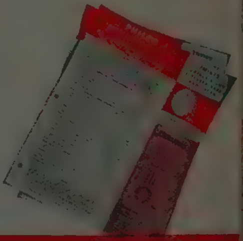


The high frequency response, together with the very low saturation voltage of these silicon Surface Alloy Diffused-base Transistors (typically 0.125 V), permits practical design of 5 mc pulse circuits, using conventional saturated switching configurations. With non-saturating techniques, pulse rates as high as 30 mc are obtainable. The typical switching circuit shown above will operate satisfactorily at trigger pulse rates up to 15 mc. When triggered with a 4-volt pulse at a 10 mc rate, the rise time will be typically less than 24 mμsec over a temperature range of −60°C to +130°C. The typical fall time will be less than 36 mμsec over the same temperature range.

Both of these transistors have demonstrated consistently more stable characteristics over a wide temperature range than any other silicon transistors available. Both meet the environmental and life test requirements of MIL-S-19500B.

NEW, MORE COMPLETE DATA SHEETS

The new data sheets on these transistors, for the first time, provide the designer with complete information upon which he may predict switching speeds in any circuit. They also contain the full military environmental and life test specifications, in accordance with MIL-S-19500B. Copies are available on request. Write Dept. IR-460.



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Third Annual Conference on Automatic Techniques

SHERATON CLEVELAND HOTEL, CLEVELAND, OHIO, APRIL 18–19, 1960

For the third consecutive year, the American Institute of Electrical Engineers, the American Society of Mechanical Engineers, and the Institute of Radio Engineers have joined together to co-sponsor the Joint Automatic Techniques Conference.

The increasing complexity in all specialized fields of endeavor makes it difficult to identify and understand the unifying theme of all modern technologies—automatic techniques. As it is envisioned, this conference draws upon the experiences of a variety of industries, obtaining both a general outline of certain problems that have confronted the industries and a discussion of the automatic techniques and the reasoning that was applied in solving the problems. Dealing broadly with techniques that have proved successful in solving practical problems, the conference appeal cuts across both industry classifications and job function categories.

April 18

Session 1—Automatic Manufacturing

"Metal Forming Under Electronic Sequencing," *E. V. Crane, Chief of Special Engineering, E. W. Bliss Co., Canton, Ohio.*

"Automatic Assembly," *J. F. Stoltz, Dept. Chief., Development Engineering, Hawthorne Works, Western Electric Co., Chicago, Ill.*

"Numerical Control Developments," *R. G. Chamberlain, Supervisor Programming, Giddings & Lewis Machine Tool. Co., Fond du Lac, Wis.*

Session 2—Automatic Inspection and Quality Assurance

"Automatic Gaging," *H. L. Boppel, Mgr. Autometrology Div., Sheffield Corp., Dayton, Ohio.*

"Production Line Testing Programmed by Punched Cards," *R. E. Wendt, Jr., Manager, Advanced Manufacturing Techniques, Headquarters Manufacturing Lab., Westinghouse Electric Corp., Pittsburgh, Pa.*

"Airframe Structural Integrity Testing," *R. L. Bondurant, Project Engineer, and D. W. Jackson, Unit Chief, Wright Air Development Center, Wright-Patterson AFB, Ohio.*

April 19

Session 3—Computer Control for Utility and Process Industries

"Operational Information System at Sterlington Steam Electric Station," *J. A. Reine, Jr., Head, Electrical Maintenance Dept., Sterlington Steam Electric Station, Sterlington, La.*

"Problems Encountered and Solved in Starting of Computer Controlled Systems," *W. F. Aiken, Project Engineering Mgr., Thompson-Ramo-Wooldridge Products Co., Los Angeles, Calif.*

"User-Supplier Relationship in Automating a Steel Rolling Mill," *C. W. Burdick, Electrical Engr., Engineering & Construction Div., Lukens Steel Co., Coatesville, Pa.; N. L. Weckstein, Equipment Engineering, Industry Control Dept., General Electric Co., Raonoke, Va.; and R. A. Hamilton, Application Engr., General Electric Co., Philadelphia, Pa.*

Session 4—Automatic Warehousing

"Automatic Baggage Handling," *H. C. Warrington, United Air Lines, Denver, Colo.*

"Automation As Applied to Book Warehousing and Shipping," *J. B. Bennett, Jr., Vice President, Secretary, and General Manager, The Macmillan Co., New York, N. Y.*

"Automatic Palletization Facilities at C and H Sugar Refining Corp.," *D. C. Gulleben, Mech. Engr., California and Hawaiian Sugar Refining Corp., Crockett, Calif.*

Tenth Symposium on Active Networks and Feedback Systems

ENGINEERING SOCIETIES BUILDING, NEW YORK, N. Y., APRIL 19–21, 1960

The tenth international symposium organized by the Polytechnic Institute of Brooklyn's Microwave Research Institute will be held on April 19–21, 1960 in the Auditorium of the Engineering Societies Building (33 W. 39 St.) New York, N. Y. This Symposium on Active Networks and Feedback Systems is cosponsored by the Professional Group on Circuit Theory of the IRE. The support of the Air Force Office of Scientific Research, the Office of Naval Research and the U. S. Army Signal Research and Development Laboratory makes it possible to hold this event without any charge for admission or registration.

This series of annual symposia has been attended by as many as 940 Scientists and Engineers in past years. Previous topics have included: Nonlinear Circuit Analysis (1953 and 1956), Modern Network Synthesis (1952 and 1955), Information Networks (1954) and four others on Millimeter Waves (1959), Microwave Techniques, Solid State Phenomena and Electronic Waveguides. As in the past, authorities will summarize the present state of the art and recent outstanding advances will be presented by research workers from university, industrial and government laboratories throughout the world.

Abstracts of all papers, a detailed program and registration form are available from The Microwave Research Institute, Polytechnic Institute of Brooklyn, 55 Johnson St., Brooklyn 1, N. Y.

Tuesday Morning April 19

The State of the Art and Future Trends

"Feedback—The History of an Idea," *H. W. Bode, Bell Telephone Labs., Inc.*

"Active Networks—Past, Present and Future," *J. G. Linvill, Stanford University.*

"Network Synthesis with Negative Resistors," *H. J. Carlin and D. C. Youla, Microwave Research Institute.*

Tuesday Afternoon

Feedback Networks and Systems

"The Methods of Feedback System Synthesis—A Survey," *J. A. Aseltine, Space Technology Labs., Inc.*

"Feedback System Evaluation," *J. G. Truxal and E. Mishkin, Polytechnic Institute of Brooklyn.*

"An Appraisal and Critique of a Class of Adaptive Systems," *I. M. Horowitz, Hughes Research Labs.*

"Feedback Systems with Non-Rational Transfer Function," *A. Papoulis, Polytechnic Institute of Brooklyn, and Institut für Hochfrequenztechnik, Germany.*

"Signal Transmission in Non-Reciprocal Systems," *J. H. Mulligan, Jr., New York University.*

"A New Design Approach for Feedback Amplifiers," *M. S. Ghausi and D. O. Pederson, University of California.*

Wednesday Morning, April 20

Synthesis of Active Networks

"The Use of the Active Lattice to Optimize Transfer Function Sensitivities," *R. E. Thomas, United States Air Force Academy.*

"Necessary and Sufficient Conditions for the Existence of ±R, C Networks," *B. K. Kinariwala, Bell Telephone Labs., Inc.*

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2 kcs to 4.9 kcs — 11/16 x 1 5/8 x 1/2
5 kcs and up — 5/8 x 15/16 x 1/2

Weight of all MLP and MHP Microids—approx. .3 ozs. each

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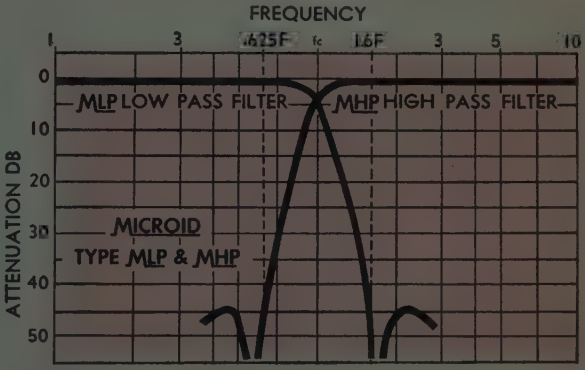

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“‘Ideal’ Three-Terminal Active Networks,” *A. W. Keen, Lanchester College of Technology, England.*

“Optimum Negative Resistance Amplifiers,” *D. C. Youla and L. I. Smilen, Microwave Research Institute.*

“Gain-Bandwidth Performance of Maximally Flat Negative-Conductance Amplifiers,” *E. W. Sard, Airbourne Instruments Lab.*

“Synthesis of Wide-Band Amplifiers,” *N. DeClaris, Cornell University.*

“High Selective Bandpass Filters Using Negative Resistances,” *M. Kawakami, Takeshi Yanagisawa and H. Shibayama, Tokyo Institute of Technology, Japan.*

Wednesday Afternoon

Active Network Concepts

“Physical Realizability of an Active Impedance,” *M. Debart, Société Alsacienne de Constructions Mécaniques, France.*

“Generalization of the Frequency-Power Formulas of Manley and Rowe,” *P. Penfield, Jr., M.I.T.*

“Bounds on Natural Frequencies of Linear Active Networks,” *C. A. Desoer and E. S. Kuh, University of California.*

“Design of Circuits to Tolerate a Range of Transistor Parameters,” *P. E. Merritt, Stanford Research Institute.*

“Polytype-Port Active Devices and Networks,” *E. E. Loebner, RCA Labs.*

“Topology and Linear Transformation Theory in Active Network Synthesis,” *E. A. Guillemin, M.I.T.*

“Formal Realizability of Linear Networks,” *Y. Oono, Kyushu University, Japan.*

Thursday Morning, April 21

Time Variable Networks

“Simulation of Orthogonal Functions,” *C. V. Jakowatz, General Electric Research Lab.*

“A Modified Form of the Mellin Transform and Its Application to the Optimum Final Value Control Problem,” *J. Peschon, Stanford Research Institute.*

“The Synthesis of Models for Time-Varying Linear Systems,” *J. B. Cruz, Jr. and M. E. Van Valkenburg, University of Illinois.*

“Periodic Solutions of Nonlinear Differential Systems,” *L. Cesari, Purdue University, and Math Res. Group, University of Wisconsin.*

“On the Properties of an Active Time-Variable Network: The Coherent Memory Filter,” *J. Capon, Federal Scientific Corp.*

“Acquisition and Tracking Behavior of Phase-Locked Loops,” *A. J. Viterbi, Jet Propulsion Lab., California Institute of Technology.*

Thursday Afternoon

Time-Varying Systems

“Linear Time Variable Transducers,” *S. Darlington, Bell Telephone Labs., Inc.*

“Perturbation Theory in Time-Varying Systems,” *A. Tonning, Norwegian Defence Research Establishment, Norway.*

“Amplification in Nonlinear Reactive Networks,” *W. R. Bennett, Bell Telephone Labs., Inc.*

Panel Discussion on Time-Varying Systems; Moderator: *S. Darlington, Bell Telephone Labs., Inc.;* Panel: *W. R. Bennett, L. Cesari, J. A. Ragazzini, A. Tonning and R. J. Schwarz.*

Twelfth Annual SWIRECO

SHAMROCK HILTON HOTEL, HOUSTON, TEXAS, APRIL 20–22, 1960

Wednesday, April 20

Session 1A—Instrumentation I

“The Responsibility of the Sales Engineer,” *R. C. McKinney, Allied Components Inc., Dallas, Tex.*

“To Buy—Or To Build—Laboratory Equipment,” *D. I. Rummer, University of Kansas, Lawrence.*

“Instrumentation of an Inertial Guidance System,” *G. W. Fuller and D. A. Scott, Jr., Chance Vought Aircraft, Dallas, Tex.*

“Analog Troque Control of Rate Sensitive Inertial Components Using a Frequency Modulated, Heterodyne Read-Out System,” *E. P. Fitzgerald, Chance Vought Aircraft, Dallas, Tex.*

Session 1B—Neurophysiology and EEG

“A Low Cost Method of Securing Integrated EMG in Real Time,” *H. M. Hanish, Litton Industries, Los Angeles, Calif.*

“A Three Channel Electromyograph with Synchronized Slow Motion Photography,” *R. W. Vreeland, et al., University of California, Medical Center, San Francisco.*

“An Active Low-Pass Filter for Biological Potential Amplifiers,” *G. White, White Instrument Lab., Austin, Tex.*

“Analysis of Neuronal Soma Action Potentials by the Method of Multiple Paired Echoes,” *R. M. Morrell, Montreal Neurological Inst., Montreal, P.Q.*

“Short Duration Hormonic Analysis of EEG Data,” *E. S. Krendel, Franklin Inst., Philadelphia, Pa.*

Welcoming Luncheon

Speaker: *Dr. R. L. McFarlan, Presiden of IRE.*

Wednesday Afternoon

Session 2A—Instrumentation II

“Automatic Testing of Radio Command Guidance Equipment,” *B. C. Hooker, Chance Vought Aircraft, Dallas, Tex.*

“A Digital Timing System for Airborne Application,” *G. P. DuBose and C. E. Grubbs, Temco Aircraft, Dallas, Tex.*

Special Bandpass Filters for Instrumentation and Control Systems,” *A. N. McGown, Jr., White Instrument Labs., Austin, Tex.*

“A Wide Band Squaring Circuit,” *R. Gaertner, Texas Instruments Inc., Dallas, Tex.*

“The Energy Necessary for Exploding Fine Wires Electrically,” *S. M. Vakil and D. L. Waidelich, University of Missouri, Columbia.*

Session 2B—Neurophysiology and EEG II

“Signal Theory Applied to the Analysis of Electroencephalograms,” *E. Lowenberg, University of Texas, Austin.*

“Period Analysis of the Electroencephalogram,” *N. R. Burch, H. E. Childers, and W. A. Spoor, Baylor University College of Medicine, Houston, Tex.*

“Averaging and Correlation Techniques Applied to Neurophysiological Measurements,” *H. W. Shipton, State University of Iowa, Iowa City.*

“An Automatic Device to Convert Electroencephalograph Records into Digital Form,” *M. G. Saunders, Winnipeg General Hospital, Winnipeg, Man.*

“Expanded Network Concepts for Analyzing Bioelectric Potentials,” *B. Saltzberg, Ramo-Wooldridge, Los Angeles, Calif.*

Wednesday Evening

Session 3A—Space Physiological Instrumentation

“Requirements of Physiological Instrumentation for Missile Flight,” *W. Greatbatch, Taber Instrument Corp., North Tonawanda, N. Y.*

“An Automatic Physiological Telemetry and Data Conversion System,” *C. A. Steinberg, W. E. Sullivan, and J. T. Farrar, Airborne Instrument Labs., Mineola, N. Y.*

Round Table Discussion on Space Medicine.

Thursday Morning, April 21

Session 4A—Computers

“Instrumentation of an Analog Seismic Correlator,” *L. C. Cummings, Jr. and F. N. Tullos, Humble Oil and Refining Company, Houston, Tex.*

“A Real Time Auto Correlator,” *S. D. Hays, Convair, San Diego, Calif.*

“Description of Rice Institute Computer,” *Dr. M. Graham, Rice Inst., Houston, Tex.*

“Engineering Details of Rice Institute Computer Memory,” *T. Schutz, Rice Inst., Houston, Tex.*

AT CAPE CANAVERAL...

Styroflex® Coaxial Cable

is a vital part of the Air Force Automatic-Tracking Antenna System!

The powerful TLM-18 telemetry antenna now in service at the Air Force Missile Center, Cape Canaveral, Fla., is used for the automatic tracking of missiles and earth satellites. This huge "mechanical ear," specifically designed by Radiation, Inc., Melbourne, Fla., has an effective data reception range of over 1000 miles.

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"Engineering Details of Rice Institute Computer Control Section," *P. Deck, Rice Inst., Houston, Tex.*

Session 4B—Biophysics I

"An Interpretation of the Effects of Atmospheric Ions on Behavior," *A. H. Fray, General Electric, Ithaca, N. Y.*

"A Non-Dispersive Infrared Gaseous CO_2 Analyzer for Physiological Studies" *L. E. Baker, Rice Inst., Houston, Tex.*

"Digital Print-Out System for Whole Body Scanner," *J. W. Beattie and G. C. Bradt, Sloan-Kittering Inst. for Cancer Research, New York, N. Y.*

"The Use of Flying Spot Scanning Techniques and Medical Research," *L. Hundley and P. O'B. Montgomery, University of Texas Southwest Medical School, Dallas, Tex.*

"An Analog Computer Model for the Study of Electrolyte and Water Flows in the Extracellular and Intracellular Fluids," *W. H. Pace, Jr., Tulane University, New Orleans, La.*

Medical Electronics Luncheon

Speaker: *Dr. H. Strughold, U. S. Air Force School of Medicine, San Antonio, Tex.*

Thursday Afternoon

Session 5A—Semiconductor Materials and Thermoelectricity

"Methods of Characterizing Semiconductor Surfaces," *C. G. Peattie, Texas Instruments Inc., Dallas, Tex.*

"Semiconductor Material Chemistry," *R. E. Johnson, Texas Instruments Inc., Dallas, Tex.*

"Fundamentals of Thermoelectric Power Generation," *T. A. Harwood, Texas Instruments Inc., Dallas, Tex.*

"Thermostating a Thermoelectric Cooler," *R. Marlow, Texas Instruments Inc., Dallas, Tex.*

"Control Circuit for Thermoelectric Devices," *A. R. Orsinger, Texas Instruments Inc., Dallas, Tex.*

Session 5B—Broadcasting

"Transistors in T. V. Receivers," *R. Webster, Texas Instruments Inc., Dallas, Tex.*

"Low Power Television Broadcast Techniques," *M. Zimmerman, Electron Corp., Dallas, Tex.*

"The Ohms-Law Phasor," *L. J. N. duTriel.*

"Directional Antennas for VHF Television Broadcasting," *H. H. Westcott, RCA, Camden, N. J.*

"Propagation Characteristics of High VHF Signals in the Immediate Vicinity of Trees," *Dr. A. H. LaGrove, University of Texas, Austin, Tex.*

Session 5C—Biophysics II

"Electronic and Computer Techniques in the Diagnosis of Cardiovascular Disease," *W. E. Tolles, W. J. Carbery and C. A. Steinberg, Airborne Instruments Lab., Mineola, N. Y.*

"Automation of Total Physiological Sensor Systems; A Reliability and Information Content Study," *B. J. Bittner and W. S. Musgrove, The Kaman Aircraft Corp., Colorado Springs, Colo.*

"Factor Analytic Methods in Medical Research," *L. W. Gaddis and W. B. Michael, University of So. California, Los Angeles.*

Periodogram and Autocorrelogram Estimates of Physiologic Circadian Periodicity by Electronic Computers—Their Use and Limitations," *F. Halberg, M.D., University of Minnesota Medical School, Minneapolis.*

"The Micro-Coulomb Method of Ozone Measurement," *G. M. Mast, Mast Development Co., Davenport, Iowa.*

Friday Morning, April 22

Session 6A—Transistor Applications

"Thermister Compensation of Transistorized Crystal Oscillators," *O. J. Baltzer, Textran Corp., Austin, Tex.*

"A Transistorized Seismic Amplifier System," *D. Venker, Texas Instruments Inc., Houston, Tex.*

"A Transistor Ferrite Core Energy Storer," *F. M. Crum, Lamar State College of Technology, Beaumont, Tex.*

"Transistor Amplifier Feed-Back Systems," *A. Parker, M.I.T., Lincoln, Lab., Lexington, Mass.*

"A Practical Approach to Transistorized Voltage Regulator Design," *E. Wilson, Texas Instruments Inc., Dallas, Tex.*

Session 6B—Physiology

"The Use of Square-Wave Electromagnetic Flow Meter to Pulse an Electrical Analogy of the Peripheral Arterial System," *M. P. Spencer, Wakeforest College, Winston-Salem, N. C.*

"Odd Problems with the Implanted Electromagnetic Blood Flow Meter: Analysis of Blood Stream Power Loss by Analog Computers and Implanted Flow Meters," *F. Olmsted, Cleveland Clinic, Cleveland, Ohio.*

"Remote Recording of Physiological Responses of Subjects Undergoing Large Amplitude Vertical Accelerations," *E. S. Gordon, Armour Research Found., Chicago, Ill.*

"Electronic Indication of Visual Fixation Point," *E. L. Thomas, Defense Research Board, Toronto, Ont.*

"Acoustic Properties of Isophageal and Artificial Larynx Speech of Laryngectomies," *H. L. Barney, Bell Telephone Lab., Murray Hill, N. J.*

Engineering Education Luncheon

Speaker: *Dr. L. Griffis, Rice Inst., Houston, Tex.*

Friday Afternoon

Session 7A—Communications

"VHF Antenna Development," *T. R. Fouts and J. I. Humphries, Chance Vought Aircraft, Dallas, Tex.*

"Regulus II Missile Program Antenna Systems," *T. R. Fouts and J. I. Humphries, Chance Vought Aircraft, Dallas, Tex.*

"A Retarded Surface Wave Antenna," *E. G. Keiffer, Chance Vought Aircraft, Dallas, Tex.*

"Sequential Detection Techniques in Pulse Communication and Radar," *M. Schwartz, Polytechnic Institute of Brooklyn, Brooklyn, N. Y.*

"System Design of the FAA Airport Surveillance Radar," *F. Simonds, Texas Instruments Inc., Dallas, Tex.*

Session 7B

Tours of medical facilities in the Houston area with inspection of specific equipment and projects.

National Aeronautical Electronics Conference

DAYTON BILTMORE AND PICK-MIAMI HOTELS, DAYTON, OHIO, MAY 2–4, 1960

The Twelfth Annual "National Aeronautical Electronics Conference" (NAECON) will be held on May 2–4, 1960 at the Dayton Biltmore and Pick-Miami Hotels, Dayton, Ohio. The theme of the Conference has been established as "Electronics Probes the Universe." In the light of recent events and those events planned for the new decade, this theme presents a timely background for the 1960 Conference.

The formulation of the papers program was highlighted by two advance actions. First, twenty session topics appropriate to the conference theme were carefully selected. Second, twenty men nationally prominent in the respective session topics were invited to serve as moderators. The session titles and moderators are listed below. These two actions were closely knit by giving each moderator the responsibility for organizing and selecting the four papers that will be presented at his session. Another feature is that several special sessions were arranged to enable "sister" professions to lend a well-

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For instance? The core has multiple isolated windings. Thus you can feed many inputs to control the amplifier. The core also has a square hysteresis loop, and thus acts as a low loss transformer. That means you save power. In addition, the core can store and remember signals—so time delay becomes simple.

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How do you want to use this superb combination? As a switching amplifier—or a linear one? In an oscillator? A power converter (d-c to d-c or d-c to a-c)? You'll have ideas of your own—and if they involve tape wound cores, why not write us? Ours are Performance-Guaranteed. *Magnetics, Inc., Dept. P-81, Butler, Pennsylvania.*

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rounded flavor to the technical program; these sessions are "Scientific Education," "Radio Astronomy," "Electric Propulsion," and "Magnetohydrodynamics."

In addition to the professional sessions, the Conference will also feature a "Student Prize-Paper Session" comprised of college student-authors from Region IV. This new generation of technical people will deliver their papers in the English Room of the Biltmore Hotel on Monday, May 2, 1960, starting at 9 A.M.

Monday Morning, May 2

Session 1—Radio Astronomy

Dr. F. T. Haddock, University of Michigan.

Session 2—Solid State Devices

Dr. I. A. Lesk, Advanced Semiconductor Lab., General Electric Co.

Session 3—Energy Conversion Systems

Dr. A. M. Zarem, President, Electro-Optical Systems, Inc.

Session 4—Ground Support

V. de Biasi, Associate Editor, Space/Aeronautics Magazine.

Monday Afternoon

Session 5—Scientific Education

Dr. R. Ewell, Vice-Chancellor for Research, University of Buffalo.

Session 6—Molecular Electronics

H. V. Noble, Technical Director, Electronic Technology Lab., WADD.

Session 7—Space Reconnaissance Systems

A. Katz, Electronic Department, Rand Corp.

Session 8—Magnetohydrodynamics

Dr. A. Kantrowitz, Director, AVCO Research Lab.

Tuesday Morning, May 3

Session 9—Space Systems Integration

D. T. McRuer, Pres., Systems Technology, Inc.

Session 10—Bionics

J. E. Keto, Chief Scientist, WADD.

Session 11—Space Communications

Dr. W. B. LaBerge, Director of Systems Management, Philco Corp.

Session 12—Circuits

C. *Wiley, President, Wiley Electronics Co.*

Tuesday Afternoon

Session 13—Interplanetary Environment

Dr. C. D. Jones, College of Engineering, Ohio State University.

Session 14—Airborne Computers

Dr. W. Seman, Computation Lab. Harvard University.

Session 15—Guidance and Control Systems

Major L. C. Wright (USAF), Bureau of Research and Development, Federal Aviation Agency.

Session 16—Antennas and Propagation

Dr. D. D. King, Vice-President, Research, Electronic Communications, Inc.

Wednesday Morning, May 4

Session 17—Managing Space Systems

J. R. Moore, Vice-President—General Manager, Autonetics.

Session 18—Microwave Tubes

Dr. L. M. Field, Director, Physics Lab., Hughes Aircraft Co.

Session 19—Telemetry

J. A. Althouse, Technical Consultant, Flight Control Lab., WADD.

Session 20—Electric Propulsion

Dr. W. J. O'Donnell, Republic Aviation

1960 Western Joint Computer Conference

JACK TAR HOTEL, SAN FRANCISCO, CALIF, MAY 3–5, 1960

The executive committee for the 1960 Western Joint Computer Conference has announced appointments to head committees preparing for the three-day conference May 3–5 at the Jack Tar Hotel in San Francisco, Calif.

Robert M. Bennett, Jr., of the IBM Research Laboratory, San Jose, Calif., general chairman, has been joined by George A. Barnard of Ampex Corp., Redwood City, Calif., vice-chairman, and J. P. Fernandez of the IBM Product Development Laboratory, San Jose, Calif., secretary-treasurer, in announcing the chairmanships.

Chairman of the technical program is Howard M. Zeidler of Stanford Research Institute, Menlo Park, Calif. Program vice-chairman is Jack E. Sherman of Lockheed Missile and Space Division, Palo Alto, Calif.

Chairman of exhibits will be Harry K. Farrar of Pacific Telephone & Telegraph Co., San Francisco, Calif.

Heading local arrangements will be Gene E. Morrison of Smith-Corona Marchant, Inc., of Oakland, Calif.

Robert A. Isaacs and Roger Wye of Philco Corp., Palo Alto, Calif., are chairman and vice-chairman for printing and mailing.

Public relations will be headed by Charles Elkind of Stanford Research Institute. Normay J. Jones of Friden, Inc., San Leandro, Calif., is public relations vice-chairman.

D. D. Willard and Earl T. Lincoln of IBM Product Development Laboratory are chairman and vice-chairman of the publications committee.

Heading the registration committee is Harold N. Wells of General Electric Computer Laboratory, Palo Alto, Calif.

Women's activities will be arranged by a committee headed by Mary Fraser of IBM Research Laboratory.

Joint sponsors of the 1960 WJCC are the Institute of Radio Engineers, the American Institute of Electrical Engineers and the Association for Computing Machinery.

Session on Computer Organization Trends

Chairman: *A. J. Critchlow, IBM, Mohansic, N. Y.*

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No man-made memory device has yet equalled the human brain from the standpoint of combined storage capacity and random-access time.

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LFE is experienced in the design, development and manufacture of Data Storage Systems for both military and industrial applications.

Further details about LFE Storage Systems may be had by writing the Vice President of Marketing. Ask for Technical Data Digest No. 6035.

L/E LABORATORY FOR ELECTRONICS, INC., Boston 15, Massachusetts

SYSTEMS, EQUIPMENT & COMPONENTS FOR:
AIRBORNE NAVIGATION • RADAR and SURVEILLANCE • ELECTRONIC DATA PROCESSING • AUTOMATIC CONTROL • INSTRUMENTATION

"The Historical Development and Predicted State-of-the-Art of the General Purpose Digital Computer," *C. Bourne and D. Ford, Stanford Research Institute, Menlo Park, Calif.*

"The Harvest System," P. O. Herwitz and J. H. Pomerene, *IBM, Poughkeepsie, N. Y.*

"Organization of Computer Systems—The Fixed Plus Variable Structure Computer," *G. Estrin, University of California, Los Angeles, Calif.*

"Horizons in Computer Systems Design," *W. F. Bauer, Ramo-Wooldridge Corp., Los Angeles, Calif.*

Panelists: *G. Amdahl, Aeronutronics Corp., Santa Ana, Calif.; M. M. Astrahan, IBM, San Jose, Calif.; and J. W. Leas, RCA, Camden, N. J.*

Session on Data Retrieval

Chairman: *R. M. Hayes, Electrada Corp.*

"A Multi-Level File Structure for Information Processing," *L. Miller, J. Minker, W. G. Reed, and W. E. Shindle, RCA, Lancaster, Pa.*

"Symbolic Logic in Language Engineering," *H. M. Semarne, Douglas Aircraft Co., Inc., Santa Monica, Calif.*

"The Fact Compiler—A System for the Extraction, Storage, and Retrieval of Information," *C. Kellogg, Ramo-Wooldridge Corp., Los Angeles, Calif.*

Panelists: *to be announced.*

Session on Learning and Problem Solving Machines

Chairman: *T. A. Kalin, M.I.T., Cambridge, Mass.*

"Empircal Explorations of the Geometry Theorem Machine," *H. Gelernter, J. R. Hansen, and D. W. Loveland, IBM, Yorktown Heights, N. Y.*

"Recognition of Sloppy Hand-Printed Characters," *W. Doyle, M.I.T., Cambridge, Mass.*

"A Suggested Model for Information Representation in a Computer That Perceives, Learns, and Reasons," *P. H. Greene, University of Chicago, Chicago, Ill.*

Panelists: *O. Selfridge, M.I.T., Cambridge, Mass.; J. C. R. Licklider, Bolt, Beranek, and Newman, Inc., Boston, Mass.; and H. J. Bremermann, University of California, Berkeley.*

Session on Analog Equipment

Chairman: *R. M. Howe, University of Michigan, Ann Arbor.*

"Analog Time Delay System," *C. D. Hofmann and H. L. Pike, Convair Astronautics, Pomona, Calif.*

"DAFT: A Digital/Analog Function Table," *R. M. Beck and J. M. Mitchell, Packard Bell Computer Corp., Los Angeles, Calif.*

"Mathematical Applications of the Dynamic Storage Analog Computer," *J. M. Andrews, Computer Systems, Inc., New York, N. Y.*

Panelists: *to be announced.*

Session on Components and Techniques

Chairman: *R. Minnick, Stanford Research Institute, Menlo Park Calif.*

"Characteristics of a Multiple Magnetic Plane Thin Film Memory Device," *K. D. Broadbent, S. Shohara, and G. Wolfe, Jr., Hughes Aircraft Co., Los Angeles, Calif.*

"Unifluxor: A Permanent Memory Element," *A. M. Renard, Aeronutronics, Van Nuys, Calif.; and W. J. Neumann, Remington Rand Univac, St. Paul, Minn.*

"A Rapid Access Transistor Driven Non-Destructive Read-Out Memory," *D. G. Fischer and T. C. Penn, Texas Instruments, Inc., Dallas, Texas.*

Panelists: *D. A. Meier, National Cash Register, Hawthorne, Calif.; and N. L. Kreuder, Electrodata, Pasadena, Calif.*

Session on Analog Techniques

Chairman: *H. Skramstad, National Bureau of Standards, Washington, D. C.*

"On the Reduction of Error in Certain Analog Computer Calculations by the Use of Constraint Equations," *R. M. Turner, Lockheed Aircraft Corp., Burbank, Calif.*

"The Use of Parameter Influence Coefficients in Computer Analysis of Dynamic Systems," *H. F. Meissinger, Hughes Aircraft Co., Los Angeles, Calif.*

"Analog Computer Techniques for Plotting Bode and Nyquist Diagrams," *G. A. Bekey and L. W. Neustadt, Space Technology Labs., Inc., Los Angeles, Calif.*

Panelists: *to be announced.*

Session on Trends in Computer Applications

Chairman: *T. W. Wilder, Broadview Research Corp.*

"Trends in Computer Applications," *J. M. Salzer, Ramo-Wooldridge Corp., Canoga Park, Calif.*

"The Outlook for Machine Translation," *F. L. Alt, National Bureau of Standards.*

"Computers for Field Artillery," *L. R. van de Velde, Fort Sill, Okla.*

Panelists: *C. E. Miller, Electronic Computing Center, San Francisco, Calif.; A. R. Zipf, Bank of America, San Francisco, Calif.; and G. W. Evans, Stanford Research Institute, Menlo Park, Calif.*

Session on Logical Design

Chairman: *R. I. Tanaka, Lockheed Aircraft Corp., Burbank, Calif.*

"Communications Within a Polymorphic Intellectronic System," *G. P. West and R. T. Koerner, Ramo-Wooldridge Corp., Los Angeles, Calif.*

"Encoding of Incompletely Specified Boolean Matrices," *T. A. Dolotta and E. J. McCluskey, Jr., Princeton University, Princeton, N. J.*

"A Built-in Table Lookup Arithmetic Unit," *R. C. Jackson, W. H. Rhodes, Jr., W. D. Winger, and J. G. Brenza, IBM Corp., Poughkeepsie, N. Y.*

Panelists: *A. Jennings, California Computer Products, Downey, Calif.; D. Aufenkamp, General Electric Co., and B. Elspas, Stanford Research Institute, Menlo Park Calif.*

Session on Design, Programming and Sociological Implications of Microelectronics

Chairman: *L. Fein, Consultant, Palo Alto, Calif.*

"On Microelectronic Components, Interconnections, and System Fabrication," *K. R. Shoulders, Stanford Research Institute, Menlo Park, Calif.*

"On Iterative Circuit Computers Constructed of Microelectronic Components and Systems," *J. H. Holland, University of Michigan, Ann Arbor.*

"On Programming a Highly Parallel Machine to be an Intelligent Technician," *A. Newell, The Rand Corp., Santa Monica, Calif.*

"On a Potential Customer for an Intelligent Technician," *C. W. Churchman, University of California, Berkeley.*

Session on Analog Applications

Chairman: *L. Wadel, Chance-Vought Aircraft Corp., Dallas, Tex.*

"ANATRAN—First Step in Breeding the DIGINALOG," *L. A. Ohlinger, Norair Division of Northrop Corp.*

"Using an Analog Computer for Both Systems Analysis and Operator Training on the Enrico Fermi Nuclear Power Plant," *R. Kley, Holley Carburetor Co.*

"Real-Time Automobile Ride Simulation," *R. H. Kohr, General Motors Corp.*

Panelists: *to be announced.*

Session on Programming Systems

Chairman: *G. H. Mealy, Rand Corp., Santa Monica, Calif.*

"A New Approach to the Programming Problem," *W. Orchard-Hays, Corporation for Economic and Industrial Research.*

"The Operator Oriented Language—A Computer Language," *G. F. Ryckman, General Motors Corp.*

"A Man-to-Machine Communication and Automatic Code Translation," *A. W. Holt and W. J. Turanski, Moore School of Electrical Engineering, University of Pennsylvania, Philadelphia (on leave from Remington Rand Univac).*

Panelists: *to be announced.*

Session on Input-Output and Communications

Chairman: *J. A. McLaughlin, IBM, San Jose, Calif.*

"A Line-Drawing Pattern Recognizer," *L. D. Harmon, Bell Telephone Labs., Murray Hill, N. J.*

"Automatic Store and Forward Message Switching System," *J. L. Owings, T. L. Genetta, and J. F. Page, RCA, Camden, N. J.*

"The Videograph Label Printing System Developed for Time, Inc.," *B. H. Klyce, Time, Inc., New York, N. Y., and J. J. Stone, A. B. Dick Co.*

Panelists: *J. Svigals, IBM, Los Angeles, Calif.; and G. Warfel, Bank of America, San Francisco, Calif.*

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DIFFUSED SILICON DIODES FROM FAIRCHILD

THE FIRST — An ultra-fast computer diode: Four millimicrosecond maximum reverse recovery time of this new FD 100 overcomes the diode-caused speed limitations in computer circuits. Capacitance is only 2$\mu\mu$f at zero volts bias.

THE REASON — A need and the technology to serve it: Fairchild's diffused silicon transistors have achieved heretofore unattainable performance. Application of these transistors has in turn created the need for silicon diodes of similarly outstanding performance.

THE FOLLOW UP — A broad line of high reliability diodes: This Fairchild FD 100 diode is being followed by others providing industry-leading standards in reliability and uniformity — backed by a continuing accumulation of statistical data on a large scale.

TENTATIVE SPECIFICATIONS — FAIRCHILD FD 100
25°C Except As Noted

Symbol	Characteristic	Min.	Max.	Conditions
BV	Breakdown Voltage	40 volts		@ I_R =100 μA
I_R	Reverse Current		.100 μA	@ V_R =30v, 25°C
V_F	Forward Voltage Drop		1 v	@ I_F =10 mA
C	Capacitance		2 $\mu\mu$f	@ V_R =0v
t_{rr}	Reverse Recovery Time To Ir=1 ma		4 mμs	@ $I_f = I_r$ =10 ma
	Maximum Power Dissipation		200 mw.	
	Temp. Range Operating	–65°C to 175°C		
	Storage	–65°C to 200°C		

For full specifications, write Dept. E-4

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PGMTT National Symposium

HOTEL DEL CORONADO, SAN DIEGO, CALIF., MAY 9–11, 1960

Welcome and Opening Remarks
Monday Morning, May 9

D. Proctor, Chairman of the 1960 PGMTT National Symposium.

A. A. Oliner, National Chairman of the PGMTT.

Session 1—Microwave Components and Systems

Chairman: *T. N. Anderson, FXR Inc., Long Island City, N.J.*

"New Developments in Microwave Communications Systems," *C. C. Cutler, Bell Telephone Labs., Holmdel, N. J.*

"Mode Conversion in Helix Waveguide," *H.G. Unger, Bell Telephone Labs., Holmdel, N. J.*

"Some Properties of Dielectric Loaded Slow Wave Structures," *G. B. Walker and C. G. Engelfield, Univ. of British Columbia, Vancouver, B. C., Canada.*

"UHF Strip Transmission Line Hybrid Junction," *I. Tatsughuchi, Bell Telephone Labs., Whippany, N. J.*

"A Variable Slope, Non-Dispersive Microwave Phase Shifter," *G. D. Carey and R. E. Hovda, Autonetics, Div. of North American Aviation, Downey, Calif.*

"A Wide-Band Turnstile Junction and Direction Finding Antenna," *R. C. Honey and J. K. Shimizu, Stanford Res. Inst., Menlo Park, Calif.*

"Microwave Phase Analyser," *K. D. Claborn and R. E. Jones, Bendix Radio, Div. of Bendix Aviation Corp., Baltimore, Md.*

Monday Afternoon
Session 2—Parametric Amplifiers

Chairman: *A. Berk, Research Labs., Hughes Aircraft Co., Culver City, Calif.*

"Gallium Arsenic Point Contact Diodes," *W. M. Sharpless, Bell Telephone Labs., Holmdel, N. J.*

"Characterization of Microwave Variable Capacitance Diodes," *S. T. Eng., Development Lab., Hughes Semiconductor Div., Newport Beach, Calif.*

"A Perturbation Theory for Parametric Amplifiers," *R. H. Kingston and A. L. McWhorter, Lincoln Lab., M.I.T., Lexington, Mass.*

"A Study of the Optimum Design of Wideband Parametric Amplifiers and Up-Converters," *G. L. Matthaei, Stanford Res. Inst., Menlo Park, Calif.*

"A Low-Noise X-Band Parametric Amplifier," *R. D. Weglein, Research Labs., Hughes Aircraft Co., Culver City, Calif., and F. Keywell, Semiconductor Div., Hughes Products, Costa Mesa, Calif.*

"A Four Terminal Low Noise Parametric Microwave Amplifier," *W. Eckhardt and F. Sterzer, RCA, Princeton, N. J.*

"Design and Operation of Four-Frequency Parametric Up-Converters," *J. A. Luksch, E. W. Matthews and G. A. VerWys, RCA, Moorestown, N. J.*

"A Study of the Iterated Traveling-Wave Parametric Amplifier," *C. V. Bell, Walla Walla College, College, Park, Wash.*

Monday Evening
Session 2A—Parametric Amplifiers

Panel Discussion: *A. Berk, Research Labs., Hughes Aircraft Co., Culver City, Calif.; C. J. Carter, Space Technology Labs., Los Angeles, Calif.; J. C. Green, Airborne Instruments Lab., Melville, L.I., N. Y.; E. M. T. Jones, Stanford Res. Inst., Menlo Park, Calif.;* N. *Uenohara, Bell Telephone Labs., Murray Hill, N. J., G. Wade, Stanford Univ., Stanford, Calif.*

Tuesday Morning, May 10
Session 3—Ferrites

Chairman: *N. Sakiotis, Motorola, Inc., Scottsdale, Ariz.*

"Non-Linear Effects in Ferrites," *M. Weiss, Research Labs., Hughes Aircraft Co., Culver City, Calif.*

"Ferroxplane Type Materials at Microwave Frequencies," *I. Bady, U. S. Army Signal Res. and Dev. Lab., Fort Monmouth, N. J.*

"Antiferromagnetic Materials for Millimeter and Sub-millimeter Devices," *G. S. Heller and J J. Stickler, Lincoln Lab., M.I.T., Lexington, Mass.*

"Design Problems Associated with Rectangular Waveguide Reciprocal Phase-shifters," *A. Clavin, RANTEC Corp, Calabasas, Calif.*

"Magnetstatic Modes and Ferrites with Conventional Waveguide Geometries," *G. P. Rodrigue, Sperry Microwave Electronics Co., Clearwater, Fla.*

"Wide Band Resonance Isolator," *W. W. Anderson and M. E. Hines, Bell Telephone Labs., Murray Hill, N. J.*

"An Electrically-Variable, Reciprocal Ferrite Phase Shifter," *R. W. Kordos and V. J. McHenry, Research Labs. Div., Bendix Aviation Corp., Detroit, Mich.*

Tuesday Afternoon
Session 4—Millimeter Waves and Diode Applications

Chairman: *P. Vartanian, Melabs, Palo Alto, Calif.*

"Millimeter Wave Generation," *B. Epsztein, Microwave Res. Inst., Brooklyn, N. Y. (on leave from Compagnie General TSF, France)*

"Millimeter Wave Generation by Parametric Methods," *G. H. Heilmeier, RCA Labs., David Sarnoff Res. Center, Princeton, N. J.*

"A Pulsed Ferrimagnetic Microwave Generator," *B. J. Elliott, T. Schaug-Petersen and H. J. Shaw, Stanford Univ., Stanford, Calif.*

"Maser Operation at Signal Frequencies Higher Than the Pump Frequency," *F. R. Ames, Airborne Instruments Lab., Melville, L. I., N. Y.*

"Generation of Microwaves by Means of Esaki Diodes," *R. F. Rutz, Research Labs., IBM, Poughkeepsie, N. Y.*

"The Large Signal Behavior of a Cavity Type Parametric Amplifier," *F. A. Olson and G. Wade, Stanford Univ., Stanford, Calif.*

"A Solid State Microwave Source from Reactance Diode Harmonic Generators," *T. M. Hyltin and K. L. Kotzebue, Texas Instruments, Dallas, Tex.*

Wednesday Morning, May 11
Session 5—Microwave Propagation in Plasmas and Solids

Chairman: *L. Goldstein, Univ. of Illinois, Urbana.*

"Propagation of Waves in a Plasma in a Magnetic Field," *W. Allis, Research Lab. of Electronics, M.I.T., Cambridge, Mass.*

"Plasma-Electromagnetic Interaction," *N. Marcuvitz, Microwave Res. Inst., Polytechnic Institute of Brooklyn, Brooklyn, N. Y.*

"Magnetoplasma Effects in Semiconductors," *B. Lax, Lincoln Labs., M.I.T., Lexington, Mass.*

"Coherent Excitation of Plasma Oscillations in Solids," *D. Pines, Univ. of Illinois, Urbana (work done at General Dynamics Corp., San Diego, Calif.)*

Panel Discussion: *O. T. Fundingsland, Raytheon Mfg. Co., Waltham, Mass.; R. Gould, California Institute of Technology, Pasadena; H. Margenau, Yale Univ., New Haven, Conn.; R. F. Whitmer, Sylvania Electronic System, Mountain View, Calif.*

Wednesday Afternoon
Session 6—Filters and Measurements

Chairman: *K. Tomiyasu, General Electric Microwave Lab., Palo Alto, Calif.*

"Lossy Waveguide Resonators," *H. Riblet, Microwave Development Labs., Inc., Wellesley, Mass.*

"Peak Internal Fields in Direct Coupled Cavity Filters," *L. Young, Westinghouse Electric Corp., Baltimore, Md.*

"Strip Transmission Line Directional Filter," *R. L. Sleven and P.E. Dorato, Airborne Instruments Lab., L. I., N. Y.*

"Application of the Smith Chart to Problems of Propagation in a Magneto-Ionic Medium," *G. A. Deschamps and W. L. Weeks, Univ. of Illinois, Urbana.*

"Microwave Measurements of Electron Attachment Rates," *V. A. J. van Lint and E. G. Wilkner, General Atomic, Div. of General Dynamics Corp., San Diego, Calif.*

"Fractional Millimicrosecond Electrical Stroboscope," *W. M. Goodall and A. F. Dietrich, Bell Telephone Labs., Holmdel, N. J.*

The banquet speaker at the PGMTT National Symposium will be W. E. Edson of the General Electric Co., Palo Alto, Calif., who will speak on "Microwave Power Sources of the Future."

Oui, General Transistor offre des transistors MIL/SPEC

GERMANIUM PNP

2N43A	MIL T 19500/18
2N44A	MIL T 19500/6
2N331	MIL T 19500/4A
2N404	MIL T 19500/20
2N416	MIL T 19500/56
2N417	MIL T 19500/57
2N425	MIL T 19500/45
2N426	MIL T 19500/42
2N427	MIL T 19500/43
2N428	MIL T 19500/44
2N464	MIL T 19500/49
2N465	MIL T 19500/50
2N466	MIL T 19500/51
2N467	MIL T 19500/45

GERMANIUM NPN

2N358A	MIL T 19500/63
2N388	MIL T 19500/65
2N1310	Guidance

SILICON PNP

2N327A	
2N328A	Guidance
2N329A	Guidance
2N1026	Guidance

DIODES

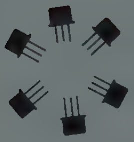
General Transistor also produces high-reliability gold-bonded diodes, three of which are designed to meet MIL requirements: 1N198, 1N277, and 1N281. The spec numbers are, respectively: MIL E 1/700, 1/993A, and 1/961.

The semiconductors listed on this page are all designed to meet MIL specs.

GENERAL TRANSISTOR CORP.

91-27 138th Place / Jamaica 35, New York

These NPN Germanium Types



As witness the deployment of General Transistors on projects such as:

BMEWS

MINUTEMAN

POLARIS

HAWK

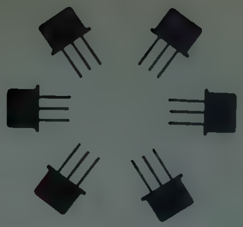
And many other military programs. Why? *Reliability.*

General Transistor is now in its fourth year of production on *quality* NPN germanium transistors. We were one of the earliest suppliers of germanium NPN transistors, and have from the beginning maintained an excellent reputation for highly reliable products. Listed are some characteristically fine types.

Type No.	Collector-Base Breakdown V_{CBO} (Volts)	Punch thru V_{pt} (Volts)	Collector Cutoff Current I_{CO} (μa)	D.C. Current Gain h_{FE}	Alpha Cutoff Frequency f_{ab} (mc)
2N356A	40	40	3	35	3
2N357A	40	40	3	40	6
2N439	40	35	3	60	8
2N440	40	30	3	100	12
2N446A	30	35	2	100	8
2N447A	30	25	2	150	12
2N595	35	30	2	50	5
2N596	35	30	2	70	8
2N1012	50	50	2	60	5

Typical values

Also on Military Duty



...and here's a flight of high voltage types:

		2N1310			2N1311			2N1312		
DISSIPATION RATINGS:		NPN GE Alloy Junction			NPN GE Alloy Junction			NPN GE Alloy Junction		
TOTAL TRANSISTOR DISSIPATION AT 25°C		120 MW			120 MW			120 MW		
DERATING FACTOR		2 MW/°C			2 MW/°C			2 MW/°C		
STORAGE TEMPERATURE		—65°C to 85°C			—65°C to 85°C			—65°C to 85°C		
CUT-OFF RATINGS:	**CONDITIONS**	**MIN**	**TYP**	**MAX**	**MIN**	**TYP**	**MAX**	**MIN**	**TYP**	**MAX**
COLLECTOR BASE VOLTAGE (V_{CBO})	I_C=25 μa	90 v			75 v			50 v		
EMITTER-BASE VOLTAGE (V_{EBO})	I_E=25 μa	20 v	50 v		20 v	50 v		20 v	40 v	
COLLECTOR-EMITTER VOLTAGE (V_{PT}) (PUNCH-THRU)	I_E=25 μa	90 v			75 v			50 v		
COLLECTOR CUT-OFF CURRENT (I_{CO})	V_{CBO}=5 v		3 μa	7 μa		3 μa	7 μa			7 μa
EMITTER CUT-OFF CURRENT (I_{EO})	V_{EBO}=5 v		3 μa	7 μa		3 μa	7 μa			7 μa
D.C. AND SWITCHING RATINGS:										
D.C. CURRENT GAIN (h_{FE})	I_C=5 ma V_{CE}=0.25 v	20			15					
	I_C=20 ma V_{CE}=0.25 v					25		20	30	
D.C. BASE VOLTAGE (V_{BE})	I_C=5 ma V_{CE}=0.25 v			0.5 v			0.5 v			
	I_C=20 ma V_{CE}=0.25 v								0.3 v	
D.C. COLLECTOR VOLTAGE (V_{CE})	I_B=10 ma I_C=100 ma		0.2 v			0.2 v			0.2 v	
SMALL SIGNAL RATINGS:										
CURRENT GAIN COMMON EMITTER (h_{fe})	V_{CE}=5 v I_E=1 ma f=1 kc		35			30			40	
ALPHA CUT-OFF FREQUENCY (f_{ab})	V_{CB}=5 v I_E=1 ma		1 mc			1.5 mc			2 mc	
COLLECTOR CAPACITY (C_{ob})	I_E=1 ma V_{CB}=5 v f=1 mc		11 μμf			11 μμf			11 μμf	
INPUT IMPEDANCE (h_{ib})	V_{CB}=5 v I_E=1 ma f=1 kc		35 Ω			35 Ω			35 Ω	
REVERSE TRANSFER RATIO (h_{rb}) (X 10^{-4})	V_{CB}=5 v I_E=1 ma f=1 kc			15			15			15
OUTPUT ADMITTANCE (h_{ob})	V_{CB}=5 v I_E=1 ma f=1 kc			2 μ℧			2 μ℧			2 μ℧
NOISE FIGURE (NF)	V_{CB}=5 v I_E=1 ma f=1 kc BW=100~		10 db			10 db			10 db	

Because of the relative newness of these transistors, data is presented in detail.

Speaking of Services: GT Hi/Scope Service

100% Lot Preconditioning

Let's assume you have equipment which must undergo severe environmental conditions...be subjected to high mechanical shock and vibration. To be certain that all the transistors you intend to use will withstand this type of exposure, we will set up a preconditioning program that will test out every single unit before we ship to you.

Special Electrical Parameter Testing

Certain transistor applications are so unusual that they cannot be completely described by standard parameters. If you are in such a position, we will design a test fixture to closely approximate actual circuit performance. This procedure will provide assurance that 100% of the transistors delivered to you will perform satisfactorily.

Special Reliability Testing Programs

Must your completed systems meet a high reliability requirement? If so, you may wish special procedures to be established with regard to your reliability programs. This is another GT service. When necessary, we will build such transistors on a specially designed production line, check them exhaustively to hold tight parameter tolerances, and subject large lots to specific and unique life tests. In many cases, we have established a program so that we ship those units which have high survival probability in your application. These things we have done, and will do again, at your request. Sound helpful?

High and Low Temperature Testing

Standard transistor parameters are generally controlled at room temperature. Yet many systems must function at other ambients. If you have a problem specifying electrical parameters at room temperature in a manner that will be valid at high or low temperatures, we are ready to assist. General Transistor is prepared to run any measurements you dictate, at any specified ambient. We can do this on complete production lots if you feel it essential.

Cost Economies Through Parameter Modifications

Yield has a strong influence on transistor cost. To give you the best economies and at the same time give you the most desirable quality, we offer this working arrangement. At your request, General Transistor will suggest slight modifications of your specifications which will allow us to ship the major portion of a production run. We will make the necessary measurements and indicate what the various parameters should be and what proportions of the run will fall into preselected types. If you then design your system to use this production mix, you will benefit from some genuine economies.

Circuit Design

If you are starting on a new program, you may want some information on what performance you can expect from state-of-the-art circuits. We will provide you with such typical circuits at your request, together with data on the performance of our transistor types within these circuits.

Special Selection on Standard Catalog Types

In many instances you may find that a standard catalog transistor is about 90% acceptable, but still needs improvement in a few parameters. In such a case, please ask us about the possibility of getting these improvements. We can tell you what increase in specifications is feasible, and produce the units to this spec. Thus, you get the desired parameters without having to redesign or wait for a custom-built semiconductor.

Qualification Approvals

Let's consider the case where you want to design a certain transistor into a system for the government, yet a government specification does not exist for the transistor. You must be ready to substantiate your use of the non-standard part. Here's what GT can do to help your case. We will run a qualification approval procedure in the same format we would for a military type. Then we'll provide you with this necessary data. This will greatly accelerate your approval for use of this transistor type.

Special Coatings or Encapsulations

In your manufacturing process, do you expose transistors to any kinds of solvents or potting materials? If so, just let us know. By using special highly resistant coatings, we'll make sure that the transistor case and markings are not vulnerable to solvent attacks.

Samples with Parameter Measurements

Assume you want to check out the margins in a design. You require upper and lower limit samples of a certain transistor type. We'll be happy to supply you with sufficient samples to cover the spread in one or two significant parameters. Thus, you can experimentally determine the performance of your circuit.

Special Production Runs

Assume that your transistor application is so unusual that units are not available from standard production. What can be done? We will analyze your requirements and decide whether it would be feasible to make a special production run of transistors to meet your needs.

What More?

This is our HI/SCOPE service...or a large part of it. Our customers have found it to be extremely useful. We think you will, too. If there is something still further that interests you, why not get in touch with us? Space precludes our going into too great detail here, but we feel sure you'll find any assistance you need at GT. Write or give us a phone call.

GENERAL TRANSISTOR CORP.

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1960 Electronic Components Conference

WASHINGTON HOTEL, WASHINGTON, D. C., MAY 10–12, 1960

Tuesday Morning, May 10

"Welcome," *R. L. Henry, Jr., General Chairman.*

Session 1—Quo Vadis, 1970

Moderator: *J. A. Chambers, Motorola.*

"Systems Requirements for Components in the Next Decade," *F. Lack, EIA.*

"The Components of 1970," *H. Stone, Bell Telephone Labs.*

"Research for Components of the Next Decade," *R. Adm. R. Bennett, Chief of Naval Research.*

Tuesday Afternoon

Session 2—Miniature and Packaged Components

Moderator: *Dr. R. M. Soria, Amphenol Electronics Co.*

"Recent Advances in Microminiaturization, Reliable Components and Cooling Techniques," *G. W. A. Dummer, Royal Radar Establishment, Great Britain.*

"Ceramic Base Microcircuits — A Heterogeneous Approach to Miniaturization," *M. H. Kahn, Sprague.*

"High Quality Ultra-Thin Film Capacitors," *S. Grand, Radiation Research Corp.*

"Rejection Filters with Distributed R&C," *A. Smith, Sprague.*

"2D Etched, Formed RC Circuits," *J. J. Kinsella and F. A. Schwertz, Haloid Xerox Inc.*

Tuesday Evening

Session 3—Special Commemorative Session—Signal Corps 100th Anniversary Presentation

Moderator: *V. Kublin, USASRDL.*

Wednesday Morning, May 11

Session 4—Microelectronic Components and Molecular Circuitry

Moderator: *C. W. Curtis, Hughes Aircraft Co.*

"The Philosophy of Molecular Electronics," *Dr. S. W. Herwald, Westinghouse.*

"Materials for Molecular Electronics," *Dr. Zener, Westinghouse.*

"Systems Aspects of Molecular Electronics," *Dr. Sziklai, Westinghouse.*

"Microelectronics and the Systems Design Engineer," *J. R. Black, Motorola.*

"Preparation and Evaluation of Thin Film Microcircuits," *J. J. Bohrer and G. N. Queen, IRC.*

"The Sylvania Microminiature Module," *G. J. Selvin, Sylvania.*

"Status Report on the Micromodule Program," *D. Mackey, J. Wentworth, RCA.*

Special Awards Luncheon

Speaker: *Maj. Gen. E. Cook, Deputy Chief Signal Officer.*

Wednesday Afternoon

Session 5—Advances in Components and Materials

Moderator: *F. Wenger, USAF-ARDC.*

"Gallium Phosphide Point Contact High Frequency Diodes," *J. Mandelkorn, USASRDL.*

"4-Layer PNPN Alloy Device," *R. Fidler, B. Carlat and R. Wegner, Tung-Sol Electric.*

"Means of Adjusting Resistive Micro Elements to the Desired Resistance Value," *R. C. Langford, Weston Instruments.*

"A Germanium Tunnel Diode," *T. R. Selig, GE, Syracuse.*

"Physics of Thin Resistive Films," *R. Conklin, Wright-Patterson AFB, E. R. Olson, Halex, Inc.*

"Ferromagnetic and Ferroelectric Ceramics; Matching Characteristics to System Needs," *Dr. H. W. Welch, Jr., Motorola.*

"Investigation of Columbium, as an Electrolytic Capacitor Anode," *C. Peabody and A. Lunchick, USASRDL.*

"Ruggedized Traveling Wave Tubes for Missile Use," *C. E. Bradford and J. P. Laico, Bell Telephone Labs.*

Thursday Morning, May 12

Session 6—Magnetic Components for Memory, Logic and Amplification

Moderator: *K. Preston, Bell Telephone Labs.*

"New Magnetic Devices for Digital Computers," *D. H. Looney, Bell Telephone Labs.*

"The Ferrite Sheet Memory," *R. H. Meinkan, Bell Telephone Labs.*

"Lesser Known Properties of Ferrite Multiapertured Cores," *U. F. Gianela, Bell Telephone Labs.*

"Reproducibility of Electrodeposited Thin Film Memory Arrays," *I. W. Wolf, GE, Syracuse.*

"Magnetic Rod Memory," *D. A. Meier, National Cash Register.*

"Counter-Wrapped Twistor," *R. F. Fische and P. Mallery, Bell, Telephone Labs.*

"The Amplifying Properties of Evaporated Magnetic Films," *D. W. Moore, Servomechanisms, Inc.*

Thursday Afternoon

Session 7—Achievements in Component Reliability

Moderator: *L. Podolsky.*

"DOD Presentation on Components Reliability Programs," *P. S. Darnell, leading a Panel Discussion.*

"Reliability Analysis of an Electronic Switching Center," *W. P. Karas, Stromberg-Carlson.*

"Electronic Parts and Reliability Assurance," *W. H. von Alven, ARINC Research Corp.*

"Reliability Analysis of Resistor Drift Characteristics," *B. R. Schwartz, RCA.*

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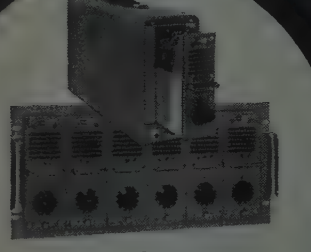

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Industrial Engineering Notes*

ASSOCIATION ACTIVITIES

Dr. Lawrence W. Von Tersch, head of the Electrical Engineering Department at Michigan State University, has been elected President of the National Electronics Conference scheduled for October 10–12 at the Hotel Sherman in Chicago. Other officers named for the 16th annual NEC, in which EIA will participate, were: Executive Vice President, Joseph J. Gershon of DeVry Technical Institute; Secretary, James H. Kogen of GPE Controls, Inc., Treasurer, Dr. Harold E. Ellithorn of University of Notre Dame; Assistant Treasurer, Robert J. Parent of University of Wisconsin. Elected as NEC Chairman of the Board was William O. Swinyard, Vice President of Hazeltine Research, Inc. Committee chairmen elected were: Arrangements, Benjamin G. Griffith, Teletype Corporation; Exhibits, John S. Powers, Bell and Howell Company; Fellowship Award, Orville I. Thompson, DeVry Technical Institute; Finance Policy, Dr. John D. Ryder, Michigan State University; Housing, Juergen Roedel, Hallicrafters Company and International Activities, George E. Anner, University of Illinois.

FCC ACTIONS

The FCC has declined to establish a TASO-type organization to complete the technical studies of the National Stereophonic Radio Committee as requested by EIA on October 15. In a letter to President Hull the FCC stated it wished to proceed expeditiously with FM stereo and that it proposed to issue a Notice of Proposed Rule Making shortly after March 15. President Hull had requested the commission to establish a Government-supervised committee similar to the Television Allocations Study Organization in response to a request from NSRC and on authorization of the EIA Board of Directors. NSRC has been unable to complete its studies, he told the FCC, because of the non-participation of RCA and CBS on legal grounds. He expressed the belief that these companies would participate if a TASO-type group were formed. Just what effect the Commission's refusal will have on the future work of NSRC was not immediately clear but will be determined within the next week or so. The NSRC was formed early in 1959 and carried on extensive technical studies under Chairman C. G. Loyd and Dr. W. R. G. Baker as Chairman of the Administrative Committee. . . . **The FCC has asked that $2,250,000 of its $13,500,000 budget request for 1961 be earmarked for the most thorough review of UHF yet.** Although plans have not been firmed up for the 2-year study, an FCC spokesman said last week the investigation would go considerably beyond previous ones on the feasibility of moving all or part of TV from VHF to UHF. Both transmitting and receiving operations would be covered. Much of the work would probably be contracted to private engineering firms, EIA learned. The commission also asked for $796,645 for research and work on frequency allocation matters. The total request increased by $2,950,000 over the 1960 appropriation.

MILITARY ELECTRONICS

The Navy has made its first public demonstration of a new communications system which uses the moon as a passive reflector or relay of radio signals. The Moon Relay included teletypewriter communications between Washington and Hawaii, via the moon, and facsimile transmissions using the same method. Admiral Arleigh A. Burke, Chief of Naval Operations said: "The new Moon Relay communications system is the result of many years of Navy development utilizing the the moon—the least expensive satellite known to man." The pilot circuit, linking Washington and Hawaii, is an outgrowth of the Navy's communication project Moon Relay, established in 1956 to develop an experimental communications system for evaluation of the moon relay technique under operational conditions. Capabilities of this experimental system presently include multi-channel teletypewriter and facsimile modes of operation. Experimental work is also being done in voice communications, but at lower priority.

SPACE LAUNCHINGS

A precise calendar of upcoming space shots was presented by Dr. Abe Silverstein, NASA's Director of Space Flight Programs. These major scientific, communications, and meteorological launchings through FY 1963 total just under 30 and include JUNO II, THOR-ABLE, SCOUT, DELTA, THOR-AGENA B, and ATLAS-AGENA B vehicles. Launchings leading up to interplanetary probes in 1963 in the vicinity of Venus and Mars, using the CENTAUR vehicle, will begin this year with three interplanetary shots using the THOR-ABLE. Next year, nine lunar launchings will be made using the DELTA, ATLAS-ABLE, and ATLAS-AGENA B, and four lunar shots will be made in 1962 employing the ATLAS-AGENA. The first launch of a payload specifically designed for the acquisition of meteorological data, known as TIROS, will be fired this summer, Dr. Silverstein detailed. This will be followed sometime in late 1961 or 1962 by

(Continued on page 40A)

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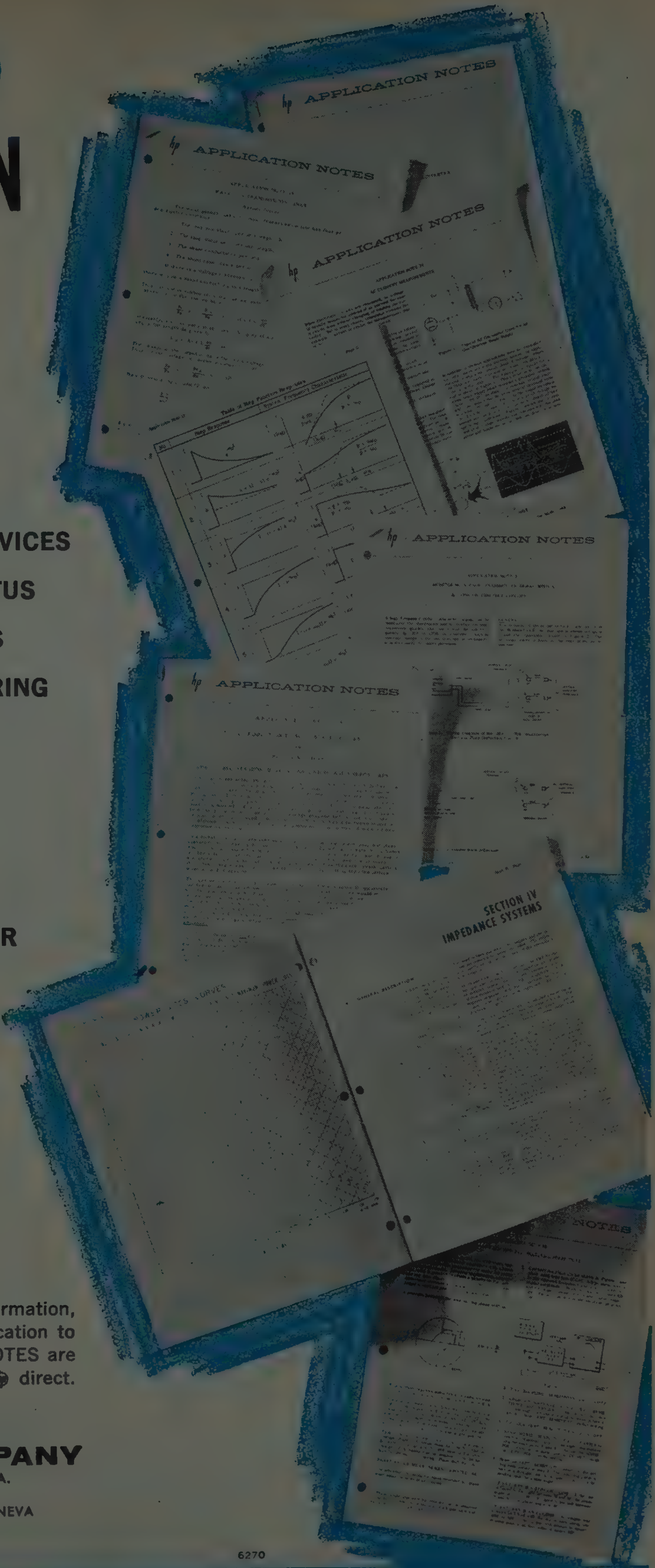
* The data on which these NOTES are based were selected by permission from *Weekly Reports*, issues of January 25 and February 1 published by the Electronic Industries Association whose helpfulness is gratefully acknowledged.

Which of these hp

APPLICATION NOTES

can help you?

- **#16** WAVES ON TRANSMISSION LINES
- **#17** SQUARE WAVE AND PULSE TESTING
- **#18** INTRODUCTION TO SOLID STATE DEVICES
- **#21** MICROWAVE STANDARDS PROSPECTUS
- **#27** BASIC MICROWAVE MEASUREMENTS
- **#29** CONVENIENT METHOD FOR MEASURING PHASE SHIFT
- **#30** MEASUREMENT OF CABLE CHARACTERISTICS
- **#34** AC CURRENT MEASUREMENTS
- **#36** SAMPLING OSCILLOGRAPHY
- **#37** MONITORING A RADIO TRANSMITTER SIGNAL WITH AN hp 120A OR 130B OSCILLOSCOPE
- **#38** MICROWAVE MEASUREMENTS FOR CALIBRATION LABORATORIES
- **#39** STANDARDS CALIBRATION PROCEDURES
- **#40** HEWLETT-PACKARD ELECTRONICS INSTRUMENTATION FOR TRANSDUCER APPLICATIONS



The above involve both theoretical and "how to do it" information, illustrated, complete, designed for swift practical application to your problem. These and all other hp APPLICATION NOTES are available by calling your hp representative, or writing hp direct. No charge, no obligation.

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EASY TO MOUNT AND STACK

Barriers across both faces increase creepage path; elongated holes facilitate mounting; nesting projection and recess aid stacking. Brass receptacles provide low contact resistance. 14 lbs. min. pull out with standard solderless taper pins. Molding compound is MAI-60 (Glass Alkyd) of MIL-M-14E.

Gen-Pro boards have passed Navy 2,000 ft. lb. high shock requirements as specified by MIL-S-901B.

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Over 25 Years of Quality Molding
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Industrial Engineering Notes

(Continued from page 38A)

a second version of the same payload, with additional sensing equipment, known as NIMBUS. This later satellite will contain more instrumentation than TIROS and will be stabilized so that the sensors will point at the earth throughout the flight path, he said. Passive communication experiments known as PROJECT ECHO—reflecting radio signals from one ground transmitting station to another—will "receive considerable attention in the months ahead," Dr. Silverstein said. But he cautioned that neither meteorological nor communications experiments "in the next several years should be considered as an early approach to an operational system." Rather, he said, these "are experiments aimed at furthering the science and technology in these areas. Operational systems will come later and only after the problems have been identified and solved." In presenting a broad picture of the national space exploration program, NASA's Deputy Administrator Hugh L. Dryden said the development of meteorological satellites is one of its most important goals. "Still in the earliest R&D stage as regards the instrumentation," he noted, "the results already obtained open new vistas to the forecaster and research scientist alike." Dr. Dryden supported the testimony of Mr. Horner.

Professional Group Meetings

Aeronautical and Navigational Electronics

Akron—December 8

"Methods of Continuous Radar Performance Monitoring," L. H. Fisher, Polytechnic Res. & Dev. Co.

Boston—January 13

"Digital Computers for Doppler Data," E. Ostroff, Lab. for Electronics, Inc.

Oklahoma City—December 22

"The FAA Air Surveillance Radar ASR-4," F. W. Simonds, Texas Instruments.

Antennas and Propagation

Boston—January 13

"A Circular-Linear Polarized Shaped-Beam Airport Surveillance Radar Antenna," A. J. Simmons and C. A. Lindberg, TRG, Inc.

"Measurements of Lunar Reflectivity Using The Millstone Radar," G. H. Pettengill, M. I. T. Lincoln Lab.

(Continued on page 44A)

Here are the plain facts!

. . . why it pays to specify and use dependable
BUSS FUSES

IT'S A FACT! By specifying BUSS fuses, you obtain the finest electrical protection possible — and you help safeguard the reputation of your product for quality and reliability.

IT'S A FACT! BUSS fuses have provided dependable electrical protection under all service conditions for over 45 years—in the home, in industry and on the farm.

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IT'S A FACT! Whatever your fuse requirements, there's a dependable BUSS or FUSETRON fuse to satisfy them. Sizes from 1/500 ampere up and there's a companion line of fuse clips, blocks and fuseholders.

IT'S A FACT! The BUSS fuse engineering staff will work with you to help you find or develop the best-suited to your needs. This places the world's largest fuse research laboratory and its personnel at your command to save you engineering time.

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BUSS fuses are made to protect - not to blow, needlessly.
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electronic, electrical, automotive and industrial use.

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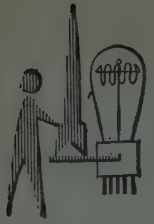
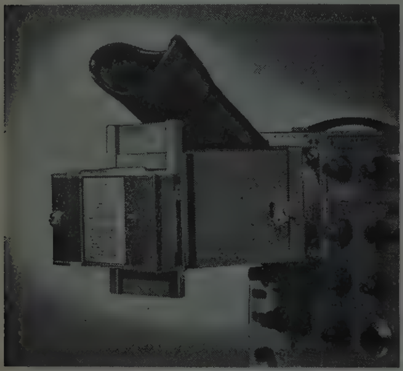


NEWS New Products

Oscilloscope Camera

A completely new oscilloscope recording camera is introduced by **Beattie-Coleman, Inc.**, 1082 N. Olive St., Anaheim, Calif. Known as the "Minute Man Oscillotron," it features a Polaroid® Land back providing either 60 second prints or transparencies. Attaches to any 5" Oscilloscope. Swings out, lifts off for accessibility. Light and compact, it is a precision instrument built for continuous duty. Wollensak 75mm f/2, 8, f/1.9 standard or f/1.9 flat-field lenses are interchangeable.

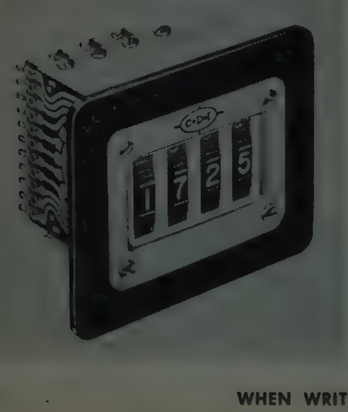


Of modular design, it can instantly be converted to record a wide range of object-to-image ratios. Can be removed from oscilloscope for other instrumentation photography.

Easily attached accessories include: Binocular viewing hood; adapter to record up to 10 traces on a single frame: 35 mm pulse camera; 35 mm continuous motion magazine; data chamber with watch, platen, and counter; data card to record in frame; external focusing control; electric shutter actuator.

Price of the basic unit is $250.00.

Digital Switch Assembly



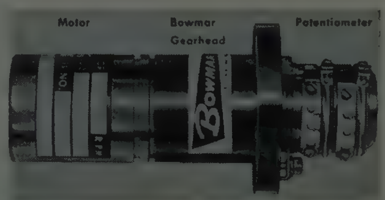
These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

A new modular digital switch assembly containing 1 to 16 thin thumbwheel switches in either 8, 10 or 12 positions with numerical readout has been designed by **Chicago Dynamic Industries, Inc., Precision Products Div.**, 1725 Diversey Blvd., Chicago 14, Ill. Series TSD may be positioned horizontally or vertically with engraved numbers or designations to suit. The unit is available with either instantly replaceable printed circuit wafers for quick servicing and low maintenance costs, or with fixed wafers if removable feature is not required. Switches are manually operated by $\frac{1}{4}$" Delrin thumbwheels available in black or colors for each module.

Immediate delivery on standard units. Prices range from $2.27 to $5.58 per module depending upon quantity and type of switch.

Miniature Female Gearhead

A new miniature female size 11 gearhead, Model X-1135, that affords easy integration of motor and potentiometers through the desired speed reduction has been developed by **Bowmar Instrument Corp.**, 8000 Bluffton Rd., Fort Wayne, Ind.



The Model X-1135 Gearhead features Bowmar's new "Postless" construction for greater shock and vibration resistance and lighter weight. A flange near the potentiometer is an integral part of the gearhead housing and is provided with three tapped holes for mounting the entire assembly into the system.

Maximum overall length, with motor and potentiometer as shown, is 3.780 inches. The gearhead with adapter and flanges is 1.250 inches long, and it weighs $1\frac{3}{4}$ ounces.

An anti-backlash arrangement may be provided on the potentiometer shaft to prevent lost motion of the unit by first assembling the gearhead to standard Bureau of Ordnance mounting holes on the size 11 motor. The potentiometer is then attached by means of standard Bureau of Ordnance type "dog" clamps.

Precision Class 2 gear tolerances per AGMA Specification 236.04 and Precision ABEC Class 5 or better ball bearings are used in the gearhead.

Additional details may be obtained from the firm.

Inexpensive Ultrasonic Cleaner

The introduction of a $99.95 ultrasonic cleaner, the diSONtegrator System Forty, by **Ultrasonic Industries, Inc.**, 141 Albertson Ave., Albertson, L.I., N.Y., will bring this equipment to a new group of users.

The System Forty, a half-gallon capacity model, is the lowest priced ultrasonic cleaner available anywhere in the world. It compares in power, cleaning tank capacity, performance and appearance with other ultrasonic cleaners now selling for up to ten times the price.

The System Forty will disintegrate more than fifty distinct classes of soils and contaminants in seconds from virtually thousands of different products ranging from surgical instruments, watches, clocks and false teeth to highly complex electronic components used in satellites and missiles.

The System Forty includes the Model G-40C1, a 40 watt generator with an output of 90,000 cps—a suitable frequency for small parts cleaning. The cabinet measures 10" L.×8" W.×$5\frac{3}{4}$" H. and features only one control knob—a simple push-pull activity regulation throttle. The diSONtegrator System Forty ultrasonic cleaner consumes no more current than an ordinary light bulb.

The generator powers the half-gallon capacity cleaning tank, Model T-4-C1, with working compartment measuring $5\frac{3}{8}$" L.×$5\frac{1}{4}$" W.×4" D. The overall dimensions are $6\frac{3}{4}$" L.×$6\frac{1}{4}$" W.×$6\frac{1}{2}$" H. The tank is heavy gauge stainless steel deep drawn with rounded corners. It is finished with a 4A grade high polish to avoid soil entrapment in crevices or corners, thereby assuring surgical cleanliness.

(Continued on page 148A)

Creative Microwave Technology

Published by MICROWAVE AND POWER TUBE DIVISION, RAYTHEON COMPANY, WALTHAM 54, MASS., Vol. 2, No. 1

RAYTHEON "M"-TYPE BACKWARD WAVE OSCILLATORS

Electronically tunable at high power levels for a wide range of microwave applications

Where extensive frequency mobility is required, the efficient crossed-field, "M"-type backward wave oscillator is highly versatile. Introduced more than eight years ago, it has been perfected by Raytheon and is now being economically mass produced. Hobbing of the slow-wave structure, a Raytheon-developed technique, assures precision construction necessary for consistently reproducible performance from tube to tube.

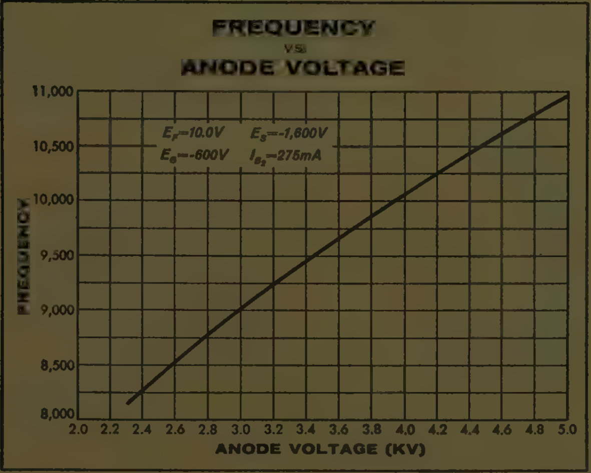
Typical of the "M"-type BWO's available from Raytheon is the QK-634A, an X-band tube which features all ceramic-and-metal construction for reliable operation under extreme environmental conditions.

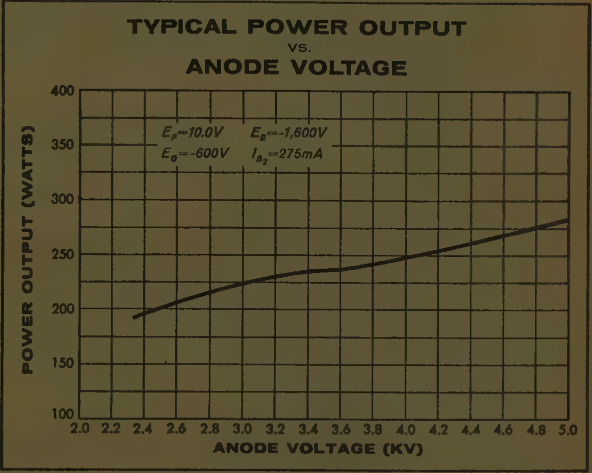
The QK-634A has a nominal power output of 200 to 250 watts and is electronically tunable over its entire frequency range. Precise determination of the radiated spectrum is accomplished by adjusting the voltage applied to either the anode or the sole. Amplitude modulation is also accomplished electronically. Small and compact, the QK-634A can be mounted in any position.

QK-634A Backward Wave Oscillator

Typical Operating Characteristics--QK-634A

Frequency Range	8,150 to 11,000 Mc
Power Output	150 watts (min.) 200 to 250 watts (nom.)
Output Flange	Mates with UG40A/U modified for clearance holes
Tuning Sensitivity	1.0 Mc/V





Other unclassified BWO's in this series include the QK-625 and QK-659, which cover the 2,500-4,450 Mc band.

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90901

INSTRUMENTATION OSCILLOSCOPE
One Inch
Miniaturized basic packaged panel mounting Cathode Ray Oscilloscope for instrumentation use replacing "Pointer Type" meters. Panel bezel matches 2" square meter. No. 90901 uses 1CP1 tube. No. 90911 uses 1EP1 tube. Power supply No. 90202 available where application requires.

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Ideal for
ANTENNA CONNECTIONS
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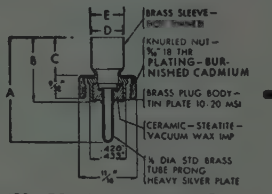
MICROPHONE CONNECTIONS



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SHIELDED TYPE
PLUGS & SOCKETS
LOW LOSS PLUGS AND SOCKETS FOR HIGH FREQUENCY CONNECTIONS

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P-202-CCT

For quality construction thruout, and fine finish, see diagram above.

101 Series furnished with ¼", .290". 5/16", ⅜", or ½" ferrule for cable entrance. Knurled nut securely fastens unit together. Plugs have ceramic insulation; sockets bakelite. Assembly meets Navy specifications.

202 Series Phosphor bronze knife-switch type socket contacts ngage both sides of flat plug contacts—double contact area. Plugs and sockets have molded bakelite insulation.

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S-202-B

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DIVISION OF UNITED-CARR FASTENER CORP.

Professional Group Meetings

(Continued from page 40A)

Chicago—October 9

"Combined Antenna-Mixer-Filter Circuit," E. M. Turner, Wright-Patterson AFB.

Chicago—December 11

"Some Investigations Performed on an Ionospheric-Scatter Test Circuit," J. D. Ahlgren, Page Communications Engineers, Inc.

Los Angeles—January 14

"Power Handling of Antennas for High Altitude Operation," T. Morita, Stanford Res. Inst.

AUDIO

Chicago—September 11

"Loudspeaker Enclosure Walls," P. W. Tappan, Warwick Mfg. Corp.

Chicago—January 8

"Sound Vibrations in a Magnetostrictive Tube," R. R. Whymark, Armour Res. Foundation.

(Continued on page 47A)

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an 18-story impact...

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Unbelievable? Not to the growing list of packaging engineers who are specifying urethane foam to protect expensive instruments, lenses, precision parts and fine tools from in-transit jars and jolts which often exceed G-forces in excess of an 18-story fall.

Urethane foam is the same material auto and aircraft makers use for shock-absorbing safety padding. Contour cut, die cut or molded in place, it is the most efficient impact-absorbing material yet invented.

The "Kudl-Pak,®" a precision part storage box used in this demonstration, is typical of the many versatile packaging ideas possible with lightweight, flexible or semi-rigid urethane foams. Cost is low; applications infinite!

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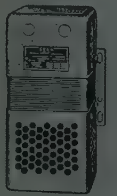
Sola Constant Voltage Transformers and Regulated DC Power Supplies provide dependable, regulated output voltage. Their output regulation is unaffected by wide variations in input voltage.

Sola CV Transformers are static-magnetic regulators with completely automatic, continuous regulating action. Their response to variations in input voltage is usually 1.5 cycles or less. They have no moving or renewable parts and require no maintenance.

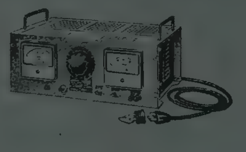
Each Sola Regulated DC Power Supply incorporates a constant voltage transformer in combination with a semi-conductor rectifier and a high-capacitance filter section. This combination makes the power supply compact, dependable, and efficient; and assures sustained output voltage in the face of pulse or intermittent loads, or heavy, short-time overloads.

Sola Constant Voltage Transformers

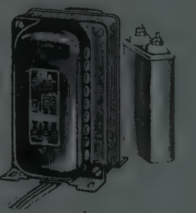
Standard Sinusoidal Type *provides voltage regulation of ±1% with primary voltage variations as great as ±15%. With less than 3% total rms harmonic content in their output voltage wave, these units are desirable for use with equipment having elements sensitive to power frequencies harmonically related to the fundamental. Available in nine ratings, 60va to 7.5kva.*



Normal-Harmonic Type *also provides ±1% regulation at somewhat less cost. This group has an average of 14% total rms harmonic content in its output voltages and is suited to equipment not extremely sensitive to voltage wave shape. The series includes those mechanical designs specially engineered for use as built-in components. Nineteen stock ratings range from 15va to 10kva.*



Adjustable Sinusoidal Type *provides ±1% regulated voltage output—one output adjustable from 0-130 volts and one fixed at 115 volts. Has less than 3% total rms harmonic content in output voltage. Portable for use in shop or laboratory, or mount on standard relay rack.*



Electronic Power Type *regulators provide ±1% regulated filament voltage at 6.0 and 6.3-volt levels; or a combination of plate and filament voltages regulated ±3% for ±15% input variations. Filament regulators are available in ratings from 2.3 to 25 amps. One model is specially designed for portable lab or shop bench use; it has a 30va rating. Combination plate/filament regulators, in three stock sizes, are designed to operate with commonly-used rectifier tubes.*

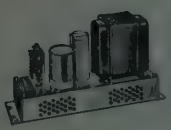
Custom-designed units *can be supplied in production quantities in ratings from 1va to 25kva to suit individual specifications. Custom designs can include special mechanical structures, various voltage ratios, special frequencies, compensation for frequency variations, multiple output voltages, three-phase service. Units can be manufactured to military specifications.*

Write for Circular 11-CV for additional information on Sola Constant Voltage Transformers

Sola Constant Voltage DC Power Supplies

For intermittent...variable ...pulse...or high-current loads

Compact, rugged Sola Constant Voltage DC Power Supplies provide output regulated within ±1% over line voltage variations as great as ±10%. Their output contains ripple voltage of 1% total rms or less. They have exceedingly low output impedance. Output is in the "ampere" range. The high, short-time overload capacity handles intermittent and pulse loads without serious voltage drop or damage to components.



Fixed-output-voltage designs *are available in six stock models with ratings from 24v @ 6a to 250v @ 1a. They are extremely compact, light-weight, and moderately priced in proportion to their power output and performance.*

Adjustable-output designs *provide a considerable range of regulated dc test voltages. Accessory handles offer portability and permit self-stacking. Six models are available with outputs ranging from 5v @ 7a to 400v @ 0.6a.*

Write for Circular 11-DC for additional information on Sola DC Power Supplies

SOLA

SOLA ELECTRIC CO.

B P
C

Sola Manufactures: Constant Voltage Transformers, Regulated DC Power Supplies, Constant Wattage Mercury Lamp Transformers and Fluorescent Lamp Ballasts

A Division of Basic Products Corporation

4633 West 16th Street, Chicago 50, Illinois, Bishop 2-1414 • In Canada, Sola Electric (Canada) Ltd., 377 Evans Avenue, Toronto 18, Ontario

Professional Group Meetings

(*Continued from page 44A*)

BROADCAST AND TELEVISION RECEIVERS

Chicago—September 11

"Cold Cathode Tubes for Commercial Receivers," A. M. Skellett, Tungsol Electric, Inc.

CIRCUIT THEORY

Chicago—October 9

"Review of Variable Reactance Amplifier Theory," L. L. Kocsis, Zenith Radio Corp.

COMMUNICATIONS SYSTEMS

Baltimore—December 8

"Statistical Communication Theory," H. Blasbalg, Electronic Communications, Inc.

Chicago—October 9

"The Sage UHF Communications Package," F. W. Griffith, II, Motorola, Inc.,

Chicago—December 11

"Some Investigations Performed on an Ionospheric-Scatter Test Circuit," J. D. Algren, Page Communications Engineers, Inc.

Los Angeles—December 8

"Digital Communications," C. Lindholm, Rand Corp.

"Digital Communications," R. Sanders, Space Electronics.

COMPONENT PARTS

Los Angeles—January 11

"A New Approach to the Organization of a Parts, Material & Process Operation," L. Jacobs, RCA.

ELECTRON DEVICES

Washington—January 18

"Status of Micro-Electronics," N. J. Doctor, Diamond Ordnance Fuze Labs.

ENGINEERING MANAGEMENT

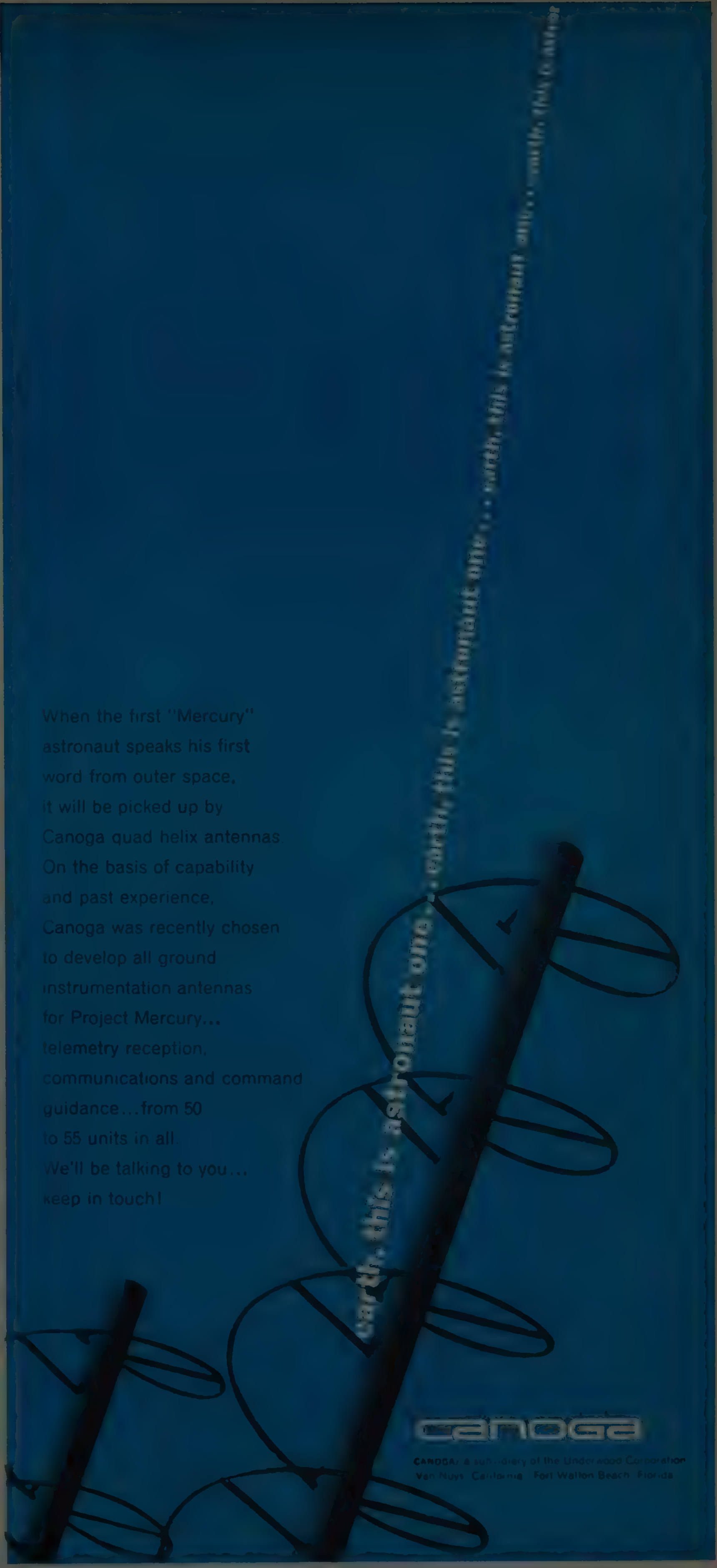
Chicago—November 13

"What Government and Industry Should Expect from Each Other," H. Neal, Cook Res.

Chicago—December 8

"The Future of the Electronics Industry," S. Weeks, Former Secretary of Commerce.

(*Continued on page 48A*)



earth, this is astronaut one . . . earth, this is astronaut one . . . earth, this is astronaut one . . . earth, this is astronaut

When the first "Mercury"
astronaut speaks his first
word from outer space,
it will be picked up by
Canoga quad helix antennas.
On the basis of capability
and past experience,
Canoga was recently chosen
to develop all ground
instrumentation antennas
for Project Mercury…
telemetry reception,
communications and command
guidance…from 50
to 55 units in all.
We'll be talking to you…
keep in touch!

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Van Nuys, California · Fort Walton Beach, Florida

BENDIX SR RACK AND PANEL CONNECTOR

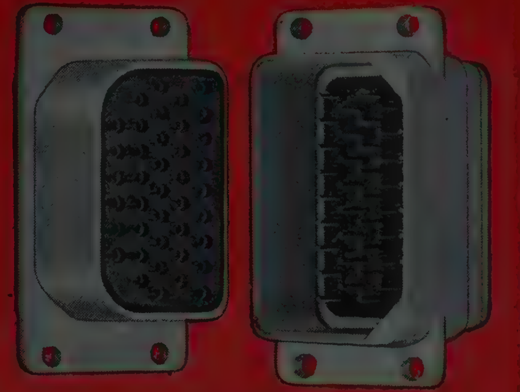
with outstanding resistance to vibration

The Bendix type SR rack and panel electrical connector provides exceptional resistance to vibration. The low engagement force gives it a decided advantage over existing connectors of this type.

Adding to the efficiency of this rack and panel connector is the performance-proven Bendix "clip-type" closed entry socket. Insert patterns are available to mate with existing equipment in the field.

Available in general duty, pressurized or potted types, each with temperature range of −67°F to +257°F.

Here, indeed, is another outstanding Bendix product that should be your first choice in rack and panel connectors.



FEATURES:

Resilient Insert • Solid Shell Construction • Low Engagement Forces • Closed Entry Sockets • Positive Contact Alignment Contacts—heavily gold plated Cadmium Plate—clear irridite finish • Easily Pressurized to latest MIL Specifications.

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SIDNEY, NEW YORK
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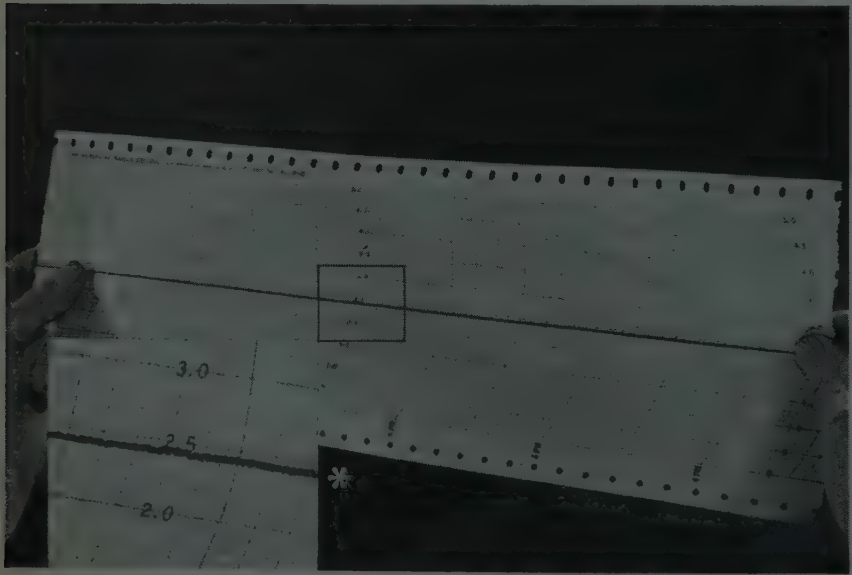
Canadian Affiliates: Aviation Electric Ltd., 200 Laurentien Blvd., Montreal 9, Quebec.
Factory Branch Offices: Burbank, Calif.; Orlando, Florida; Chicago, Illinois; Teaneck, New Jersey; Dallas, Texas; Seattle, Washington; Washington, D. C.



Measure fractions of a microvolt...approaching the Johnson noise limit...with Beckman DC Breaker Amplifiers. These high gain, low drift amplifiers are insensitive to vibrations, provide fast response and feed outputs directly to standard recorders. This means you can measure dc and low frequency ac voltages which were impossible or too tedious with devices like suspension galvanometers. A few applications include use with ultra-precision bridge circuits for measurement of differential thermocouples, nerve voltages, and other extremely low voltages. For detailed specifications write for Data File 9-4-11.

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Beckman Instruments, Inc.
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INSTRUMENTAL IN YOUR FUTURE 25 YEARS 1935-1960

*Note low noise level...less than .003 microvolt

Professional Group Meetings

(Continued from page 47A)

INSTRUMENTATION

Washington—January 11

Panel Discussion on "The Next Ten Years in Instrumentation," S. N. Alexander, NBS; R. L. Bowman, National Heart Institute of NIH; T. MacAnespie, Glen L. Martin Co.

MEDICAL ELECTRONICS

Boston—December 8

"Digital Computer Techniques in the Study of the Nervous System," W. A. Clark, Lincoln Lab. M.I.T.

Boston—January 26

"Engineering for Surgery," J. J. Baruch, Bolt, Beranek and Newman.

Chicago—December 11

"The Clinical Quantitation of Tone and Tremor," J. Brumlik, M.D.; H. Wachs, M.D.; B. Boshes, M.D.; M. L. Petrovick, B.S.E.E.; Northwestern University Medical School.

(Continued on page 50A)

The Most Complete Line of

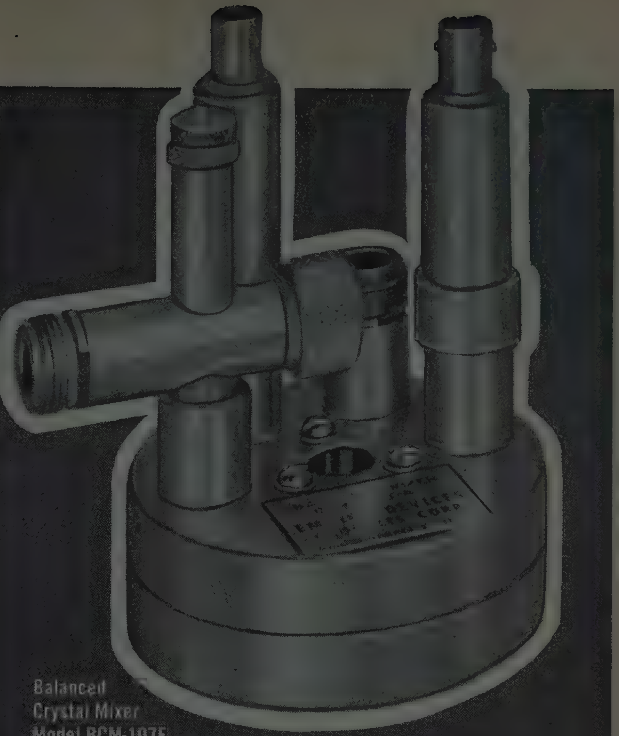
COAXIAL BROADBAND CRYSTAL MIXERS

We have the answers to many unique coaxial mixer design requirements.

• Broadbanding • Low VSWR • Unusual Frequency Combinations • Low-Noise Design • R.F.-L.O.-I.F. Isolation.

The expanded line of coaxial mixers from EMPIRE represents the solution to these, and many other problems. Most types are available for immediate delivery from stock.

WHAT IS YOUR SPECIAL PROBLEM?



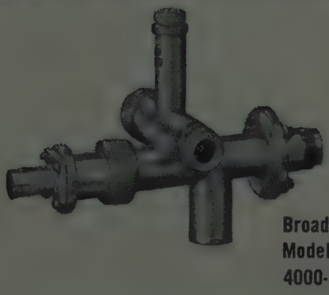
Balanced
Crystal Mixer
Model BCM-107E
2500-2900 MC.

BALANCED MIXERS, Model BCM-107

For R.F. frequencies from 1000 to 5600 MC.
Models for push-pull or single-ended I.F. inputs.
Featuring, over a 15% bandwidth:

★ Input VSWR less than 1.5 to 1, without adjustment.
★ L.O. noise cancellation at least 20 DB
★ Isolation between R.F. and L.O. at least 20 DB.
★ L.O. rejection at I.F. terminals at least 30 DB

Broadband Crystal Mixer
Model CM-107A2
510-760 MC.



Broadband Crystal Mixer
Model CM-107F
4000-5600 MC.

BROADBAND MIXERS, Model CM-107

For R.F. frequencies from 225 MC to 8200 MC.
Featuring over a broad range of frequencies:

★ Input VSWR less than 2 to 1, without adjustment.
★ L.O. VSWR less than 2 to 1, without adjustment.
★ L.O. rejection at I.F. output, better than 30 DB.
★ L.O. power requirements 10 MW, with adjustable injection feature.

For complete data, send for Catalog 604.

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AMSTERDAM, NEW YORK
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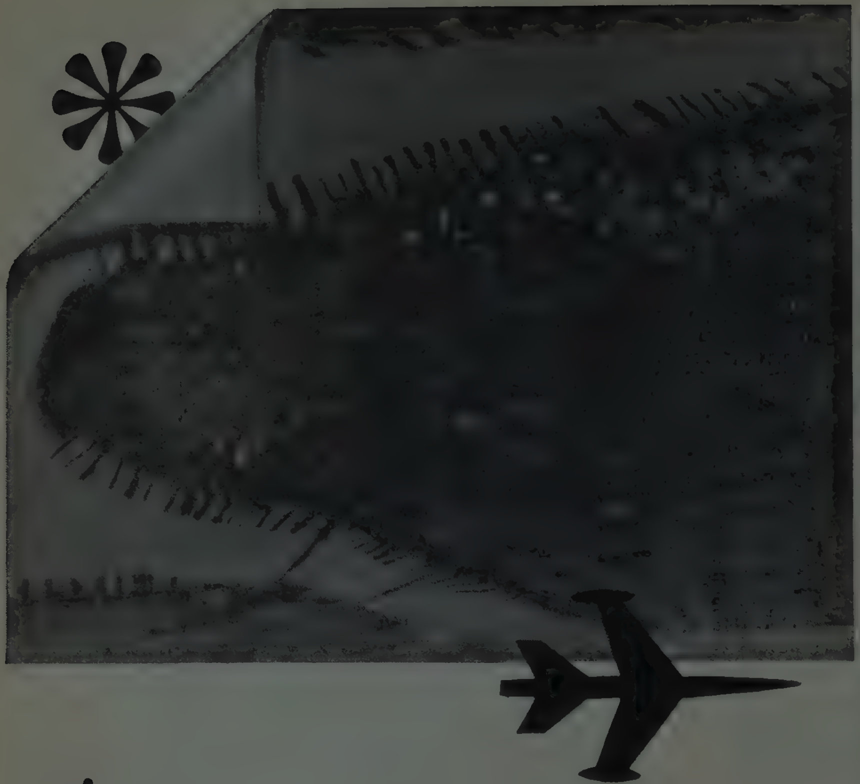
MANUFACTURERS OF:
FIELD INTENSITY METERS • DISTORTION ANALYZERS • IMPULSE GENERATORS • COAXIAL ATTENUATORS • CRYSTAL MIXERS



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IR surveillance equipment which meets military requirements, is continually being developed and improved at HRB-SINGER. Although RECONOFAX, the trade name applied to HRB IR equipment has been employed primarily in aircraft, it could be used in other vehicles such as satellites for scanning areas several hundred miles wide.

If you are interested in HRB's outstanding advances in the development of new concepts and systems for reconnaissance, surveillance, and infrared detection—military and industrial personnel with a need to know, contact HRB-SINGER, Dept. I.

ELECTRONIC RESEARCH AND DEVELOPMENT in the areas of:

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A SUBSIDIARY OF THE SINGER MANUFACTURING COMPANY
Science Park, State College, Pa.

SINGER

Professional Group Meetings

(Continued from page 48A)

MILITARY ELECTRONICS

Chicago—October 9

"The SAGE UHF Communications Package," F. W. Griffith, II, Motorola, Inc.

Chicago—November 13

"What Government and Industry Should Expect from Each Other," H. T. Neal, Cook Electric.

Indianapolis—December 10

Election of Officers.

Northwest Florida—December 15

"Recent Biological Experiments in Ballistics Missiles," D. E. Stullken, Naval Air Station.

Philadelphia—January 19

"Sonic Detection Techniques," J. Howard and H. West, Naval Air Dev. Center.

PRODUCTION TECHNIQUES

Boston—January 12

"Numerical Control and its Use in Industry Today," H. P. Kilroy, Concord Control, Inc.

Philadelphia—January 20

"Molectronics," G. Strull, Westinghouse Electric Co.

"Micro-Modules," J. Wentworth, RCA.

"Integrated Electronics," R. Lee, Texas Instruments, Inc.

SPACE ELECTRONICS AND TELEMETRY

Philadelphia—January 20

"Molectronics," G. Strull, Westinghouse.

"Micro-Modules," Mr. Wentworth, RCA.

"Micromolecular Circuits," R. Lee, Texas Instruments.

Washington—November 17

"Electrical Propulsion for Space Missions," W. Moeckel, Lewis Res. Centr. NASA.

VEHICULAR COMMUNICATIONS

Los Angeles—January 21

"MT-59 ITT Mobile Telephone Equipment Design and Operation," K. Haase, International Telephone & Telegraph Co.

Use Your
IRE DIRECTORY!
It's Valuable

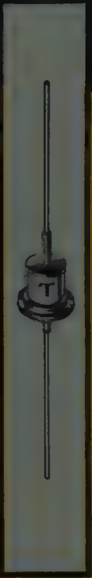
NEW . . . FROM INDUSTRY'S BROADEST RECTIFIER LINE

SILICON CARBIDE RECTIFIERS

FEATURES

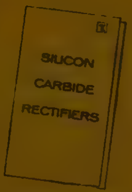
- IMPROVED RELIABILITY AT ELEVATED TEMPERATURES
- HIGHER TEMPERATURE PERFORMANCE TO 500°C
- EXCELLENT RADIATION RESISTANCE

Silicon carbide rectifiers are now available for production requirements. Their inherent higher temperature capabilities (to 500°C) provide greater reliability in existing circuitry subject to high ambient temperatures. Resistance to radiation is up to 100 times greater than previously available in any semiconductor rectifiers.



	SPECIFICATIONS @ 500°C.			RATINGS @ 500°C.	
Type	Peak Inverse Voltage (volts)	Maximum Inverse Current I_b (μa)	Maximum Forward Voltage @ Specified Current (volts @ mA)	Maximum Average Forward Current I_o (mA)	Maximum Peak Inverse Voltage (volts)
TCS10	100	500	6 @ 100	100	100
TCS5	50	500	4 @ 100	100	50

	SPECIFICATIONS @ 25°C.		
Type	Peak Inverse Voltage (volts)	Maximum Inverse Current I_b (μa)	Maximum Forward Voltage @ Specified Current (volts @ mA)
TCS10	100	10	12 @ 100
TCS5	50	10	8 @ 100



Write for Bulletin PB-70

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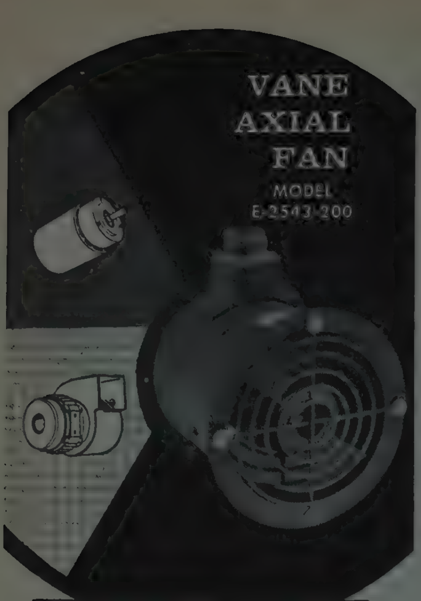

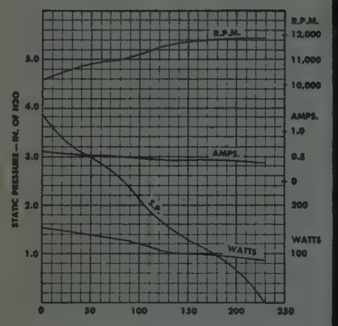

SPECIFICATIONS:
200V 400 cps 3 phase
200 CFM at 3/4" S.P.
Weight: 3 lb. 8 oz.
Ambient: 85°C
Life: 5000 hrs.
Environmental MIL-E-5422D
Material MIL-E-5400B
Class F Insulation

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WRITE TODAY FOR OUR NEW CATALOG

IRE People

The appointment of **Gordon C. Berry** (S'43–A'47) as regional sales engineer for receiving tubes has been announced by C. Byron Farmer, southeastern regional sales manager of the General Electric Receiving Tube Department.

G. C. Berry

In his new position, Mr. Berry will provide engineering liaison with both manufacturers of electronic equipment and with users of equipment in the commercial and industrial electronics field. For several years he has served as a district engineer working out of Atlanta on two-way radio, microwave, and broadcasting engineering assignments.

A native of Huntsville, Ala., he is a graduate of Georgia Institute of Technology from which he received the B.S. degree in electrical engineering in 1947. During World War II he served as a first lieutenant in the Army Signal Corps.

Mr. Berry is a member of the National Society of Professional Engineers.

❖

Glen P. Bieging (M'56) with 20 years' experience as an executive in military and industrial electronics, has joined Packard Bell Electronics Corp. as vice president, marketing, of the company's new Defense and Industrial Group, it was announced by Richard B. Leng, group vice president.

G. P. Bieging

For the last five years Mr. Bieging has been associated with General Electric as manager of product planning for the Heavy Military Electronics Department in Syracuse, N. Y., and as manager of marketing for the Missile and Space Vehicle Department, Defense Electronics Division, Philadelphia.

He served as an electronics officer in the U. S. Air Force from 1940–50, resigning with the permanent rank of major. During the Korean conflict he spent one year as civilian consultant to the Deputy Chief of Staff, Development, Headquarters USAF, then moved to Raytheon Manufacturing Co., Waltham, Mass., as assistant to the vice president for engineering and research. He subsequently served as salesmanager of Lavoie Laboratories, Morganville, N. J., and operated his own company in the electronics export field.

At Packard Bell Electronics, he will direct marketing and product planning for the Defense and Industrial Group which comprises the Technical Products Division, Packard Bell Computer Corp. and Physical Sciences Corp., a subsidiary formerly known as Technical Industries Corp.

Mr. Bieging holds a B.S. degree in business and engineering from the University of Minnesota and was graduated from the British Military College of Science. He also attended the Air Command and Staff School and the Air Communications and Electronic Staff Officers School. He is a member of the American Ordnance Association, Air Force Association, American Society of Naval Engineers and the Armed Forces Communications Electronics Association.

❖

The New York Academy of Science announced the award to the grade of Fellow to **Roland C. Bostrom** (S'49–A'50–M'55) for his work in the field of Medical Electronics.

A supervisor in the Department of Medical and Biological Physics of Airborne Instruments Laboratory, Deer Park, N. Y., he is known for his work on the Cytoanalyzer, an electronic instrument that detects cancer in the shed cells of the body by means of the Papanicolaou smear test.

The Cytoanalyzer is presently under evaluation at The National Cancer Institute of Health, Hagerstown, Md. and the Sloan Kettering Institute, New York, N. Y. This equipment has been designed to detect cancer of the uterus by scanning microscope slides bearing the specimen cells.

❖

H. Thomas Bean (M'57), Vice-President of Telecomputing Services, Inc., and General Manager of their data reduction operations at White Sands Missile Range has been elected Chairman of the Executive Committee of the Holloman Test Directors Council at the Air Force Missile Development Center, N. M.

In addition to his duties as Administrator of the data reduction facility at Holloman Air Force Base, he is chairman of the Management Committee, Data Reduction Group, IRIG; and chairman of the New Mexico Advisory Committee on Scientific, Engineering and Specialized Personnel.

Mr. Bean received the B.S. degree in physics from Thiel College. He also has been an instructor in physics at Thiel College, Chief Physicist for the 7th Naval District, Chief of Data Reduction at Eglin Air Force Base, and a member of the Technical Staff, Space Technology Labs before joining TSI in his present capacity.

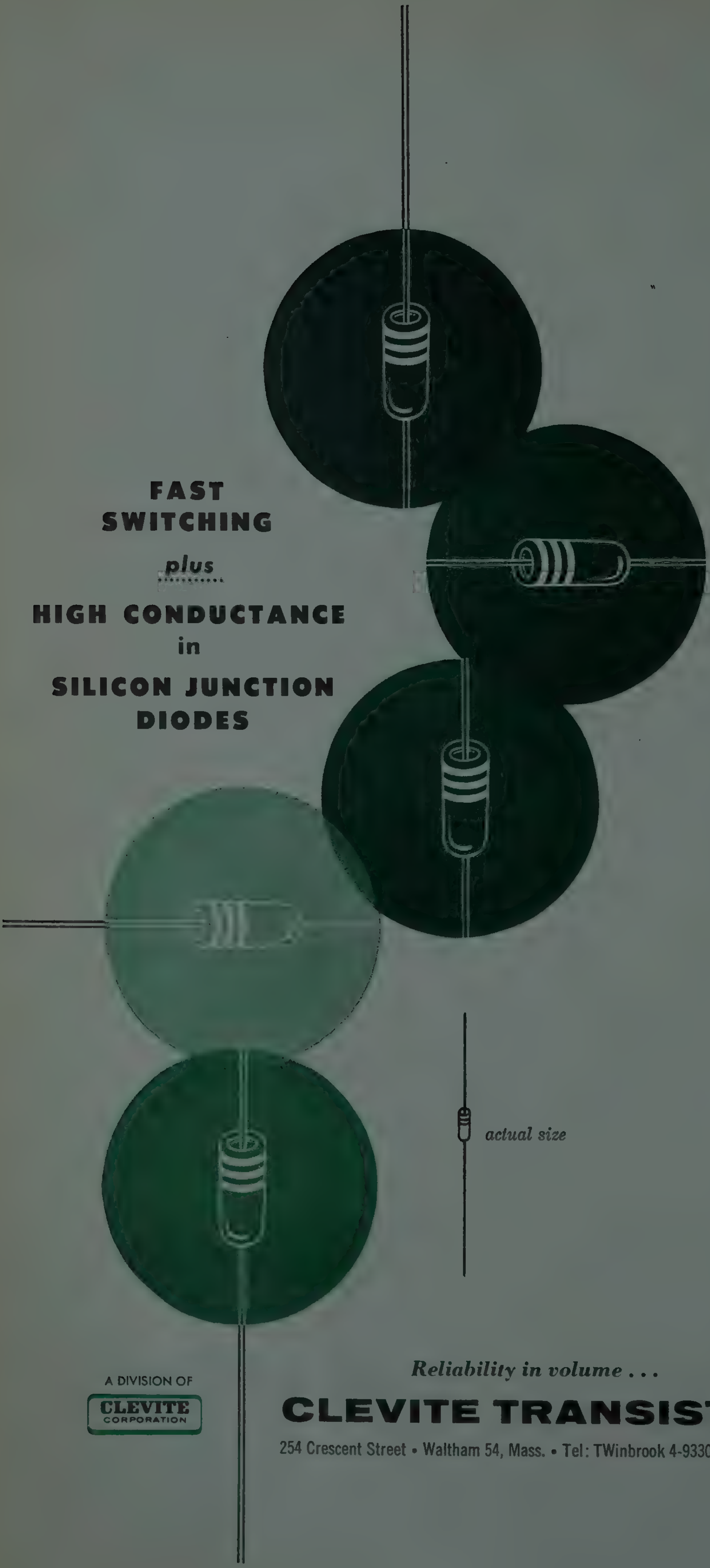
He is a member of the American Rocket Society and Instrument Society of America.

❖

(Continued on page 56A)

Reliability in volume...

CLEVITE
TRANSISTOR
WALTHAM, MASSACHUSETTS



FAST SWITCHING

plus

HIGH CONDUCTANCE

in

SILICON JUNCTION DIODES

SWITCHING TYPES

New circuit possibilities for low impedance, high current applications are opened up by Clevite's switching diodes. Type CSD-2542, for example, switches from 30 ma to −35v. in 0.5 microseconds in a modified IBM Y circuit and has a forward conductance of 100 ma minimum at 1 volt.

Combining high reverse voltage, high forward conductance, fast switching and high temperature operation, these diodes approach the ideal multi-purpose device sought by designers.

GENERAL PURPOSE TYPES

Optimum rectification efficiency rather than rate of switching has been built into these silicon diodes. They feature very high forward conductance and low reverse current. These diodes find their principal use in various instrumentation applications where the accuracy or reproduceability of performance of the circuit requires a diode of negligible reverse current. In this line of general purpose types Clevite has available, in addition to the JAN types listed below, commercial diodes of the 1N482 series.

MILITARY TYPES
JAN

1N457	**MIL-E-1/1026**
1N458	**MIL-E-1/1027**
1N459	**MIL-E-1/1028**

Signal Corps

1N662	**MIL-E-1/1139**
1N663	**MIL-E-1/1140**
1N658	**MIL-E-1/1160**
1N643	**MIL-E-1/1171**

All these diodes are available for immediate delivery. Write now for Bulletins B217A-1, B217A-2 and B217-4.

Phone for data and prices.

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- *Widest range!*

NARDA transistorized

POWER METER

BATTERY MINIMUM
POWER MILLIWATTS
the narda
microwave corporation
BATTERY CHECK
CHARGING
BIAS RANGE

MODEL 440 **. . . $250**

What's most important to you in a power meter? Accuracy? Portability? Independence from line voltage deviations? Wide range? Stability? Rapid warm-up?

Not that you have to make a choice . . . or a compromise . . . any longer. The Narda Model 440 Power Meter gives you all these features! Completely transistorized and powered by a nickel-cadmium battery, rechargeable during operation or overnight, it offers two low-power scales in addition to the five standard scales (see below), a built-in charger with state-of-charge indicator and protection against overcharging, and freedom from internal heating caused by vacuum tubes.

Moreover, the 440 provides up to 18 ma bias current, enabling you to use the widest selection of bolometers and thermistors. In short, the 440 is the most versatile unit available to provide accurate direct-reading measurements of cw or pulsed-power automatically, over any frequency range for which there are bolometer or thermistor mounts. For complete data, contact your nearest Narda representative, or write us directly. Address: Dept. PIRE-11.

SPECIFICATIONS

POWER RANGES: 7 SCALES

***0.01 mw full scale**	—30 to —20dbm
***0.03 mw full scale**	—25 to —15dbm
0.1 mw full scale	—20 to —10dbm
0.3 mw full scale	—15 to —5dbm
1.0 mw full scale	—10 to 0dbm
3.0 mw full scale	—5 to +5dbm
10 mw full scale	0 to +10dbm

*4.5 ma bolometers give best results on these scales.

Range Switch: 0.01 to 10 mw (full scale)

Accuracy: 3% of full scale reading

Bolometers & Thermistors: All 100 and 200 ohm, requiring up to 18 ma bias.

Battery Charger: Built-in; continuous or overnight. (Battery operable 16 hrs. before recharge required.)

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118-160 HERRICKS ROAD, MINEOLA, L. I., N. Y. • PIONEER 6-4650

Miniature

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For over a decade HYSOL chemists and engineers have been helping manufacturers solve unique and intricate problems of insulation. As a result of this experience, Hysol has developed a complete line of epoxy encapsulating compounds. For superior insulation, for outstanding moisture, chemical and abrasion resistance, for dependable performance . . . there's a Hysol epoxy to meet your specifications. Write for the HYSOL "Systems Selector."

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IRE People

(Continued from page 52A)

Dr. P. S. Christaldi (S'35–A'40–SM'44–F'52) has joined G-V Controls Inc. of Livingston, N. J. as Manager of Engineering. He has been, since 1956, Product Manager, Nuclear Systems, for Curtiss-Wright Corp. Prior to that he was associated for eighteen years with Allen B. DuMont Laboratories, Inc.

P. S. CHRISTALDI

His appointment represents the first move in a program of expansion and diversification planned by G-V Controls, following the company's move into a newly constructed plant in Livingston. Present products include thermal relays, electrical thermostats, transistorized time delays, and wired control assemblies.

Dr. Christaldi is past Chairman of the Northern New Jersey Section of the IRE, and the author of a number of papers in the electronic field and of the section on Oscilloscopes and Electronic Switching of the McGraw-Hill Industrial Electronics Handbook.

❖

Frank K. Clark Jr. (S'48–A'51–M'56) has been appointed chief engineer of the M. C. Jones Electronics Co., Inc., it was announced by George E. Steiner, president of the Jones company and general manager of the Scintilla division of Bendix Aviation Corporation.

The Jones company, a subsidiary of Bendix, manufactures test equipment for monitoring radio frequency coaxial transmission lines.

Clark joined the Bendix Radio division in Baltimore in 1954 and worked on the development of high-powered radio transmitters and microwave tube applications. Previously he was with Electro Precision Products, Inc. and Hazeltine Electronics Corporation.

He is a member of the American Ordnance Association. In 1958 he served as secretary-treasurer of the Baltimore section of the IRE and in 1959 was vice chairman. He is a 1950 graduate of the University of Texas with a B.S. degree in electrical engineering.

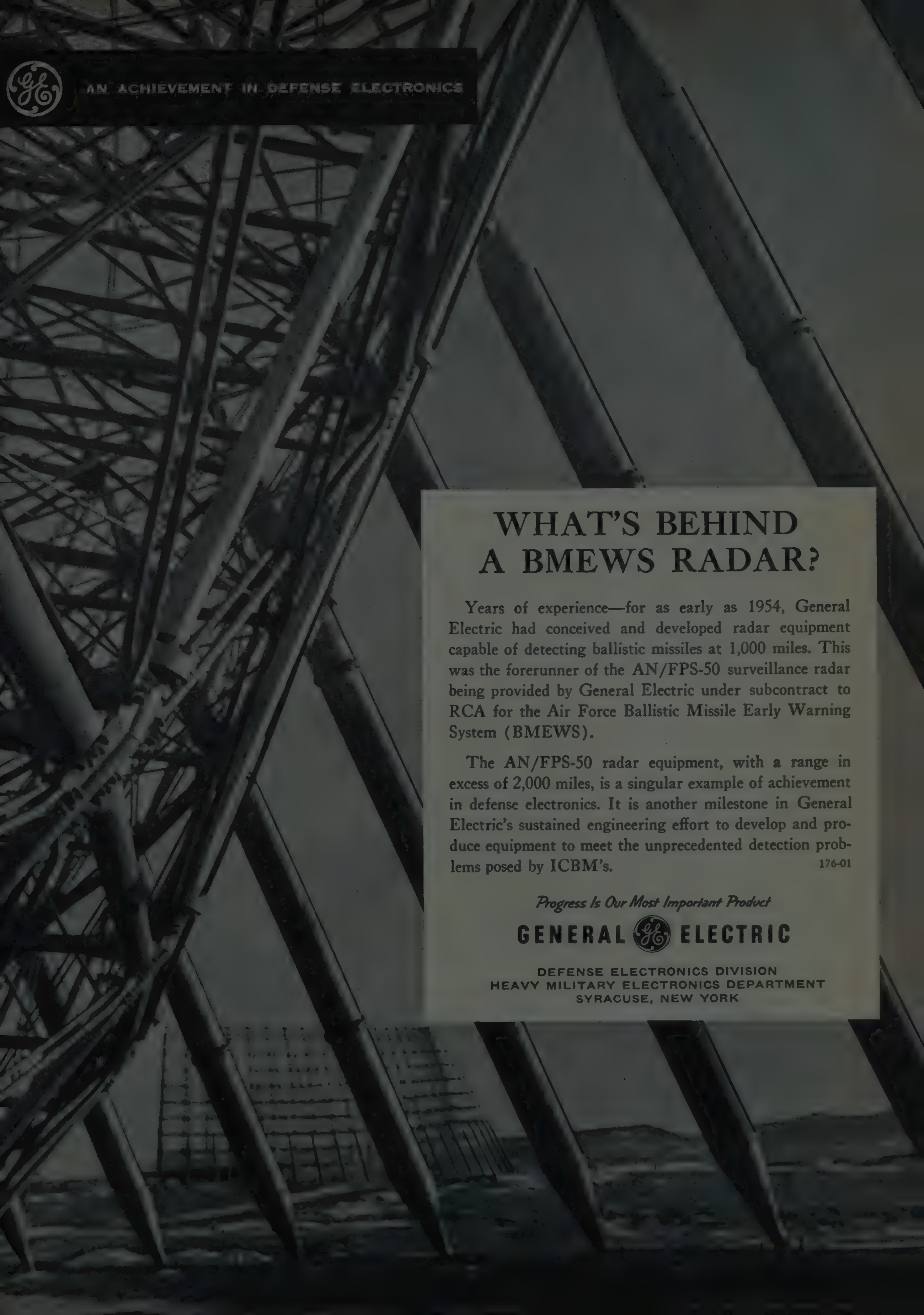
❖

The promotion of **Paul W. Crapuchettes** (M'44–F'60) to Technical Director of the Litton Industries Electron Tube Division was announced recently by Dr. Norman Moore, Division General Manager.

In addition to this Senior staff position, he is also Manager of the Magnetron Product Line.

Mr. Crapuchettes was formerly Chief Engineer for the Electron Tube Division. He has been associated with the Litton In-

(Continued on page 58A)



AN ACHIEVEMENT IN DEFENSE ELECTRONICS

WHAT'S BEHIND A BMEWS RADAR?

Years of experience—for as early as 1954, General Electric had conceived and developed radar equipment capable of detecting ballistic missiles at 1,000 miles. This was the forerunner of the AN/FPS-50 surveillance radar being provided by General Electric under subcontract to RCA for the Air Force Ballistic Missile Early Warning System (BMEWS).

The AN/FPS-50 radar equipment, with a range in excess of 2,000 miles, is a singular example of achievement in defense electronics. It is another milestone in General Electric's sustained engineering effort to develop and produce equipment to meet the unprecedented detection problems posed by ICBM's. 176-01

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DEFENSE ELECTRONICS DIVISION
HEAVY MILITARY ELECTRONICS DEPARTMENT
SYRACUSE, NEW YORK

The B-58 Hustler, America's first supersonic bomber, is now in production for the U. S. Air Force at the Fort Worth, Texas, Plant of the Convair Division of General Dynamics Corporation. Carrying a crew of only three, the B-58 is designed to operate at altitudes above 50,000 feet and is powered by four General Electric J79 turbojet engines.

Sprague Awarded Contract For Engineering Assistance On B-58 Hustler

The abundance of electronic equipment fitted into the small, dense, supersonic B-58 Hustler calls for comprehensive interference control engineering. To obtain on-the-spot assistance for this extremely complex engineering problem, the Weapon System Manager for the Mach 2 bomber at Convair's Fort Worth Plant has awarded a special contract to the Interference Control Field Service Department of the Sprague Electric Company.

Under the terms of the contract Sprague will assist Convair with the integration of both electrical and electronic equipment of airborne and ground systems into an even more effective and reliable weapon system. Sprague staff engineers assigned to the project are already at work at the Fort Worth Plant.

Sprague Interference Control Laboratories located in Dayton, Los Angeles, and North Adams, are supporting the Convair program as testing and sampling facilities. These laboratories are staffed by top interference and susceptibility control specialists, and are equipped with the most advanced instrumentation and model shop facilities.

For assistance with your Interference, Susceptibility, and Integration problems, write Interference Control Field Service Manager, Sprague Electric Co. at 12870 Panama Street, Los Angeles 66, California; 224 Leo Street, Dayton 4, Ohio; or 235 Marshall Street, North Adams, Massachusetts.

IRE People

(Continued from page 56A)

dustries Engineering Department since 1946. Mr. Crapuchettes was graduated from the University of California in 1940.

❖

Alan F. Culbertson (A'55–SM'56) has been appointed Director of Engineering at Lenkurt Electric Co., Inc., San Carlos, Calif.

A. F. Culbertson

He will be responsible for development of telecommunications and allied systems for commercial and government use, as well as for supporting engineering activities including component development and quality control.

He has been with Lenkurt, a subsidiary of General Telephone and Electronics, since 1952 when he was employed as an applications engineer. He later was manager of transmission engineering and of product planning. For the past year and a half he has headed the mobile telephone group which has developed and marketed new systems for vehicular telephone service.

(Continued on page 60A)

KRYSTINEL FERRITES

K-202 FERRITE CORES

Now in production quantities, this ferrite material designed especially for use as choke cores, tuning cores, threaded cores, cup cores, in 30 kc. to 3 mc. frequency range.

- excellent magnetic properties
- stable shock and aging characteristics
- low permeability drift with respect to temperature . . . consistently uniform

Krystinel offers a comprehensive line of ferrite materials, covering a wide range of electronic applications and frequencies. Write for Technical Bulletin Series "F".

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Delay lines at ESC are now scheduled, produced and inspected under the control of a completely automated, electronic IBM Integrated Data Processing System. The new system enables ESC to know, within minutes, the status of every delay line order. Vital delivery information can now be presented with greater precision. Statistics, now immediately available on production runs, serve as invaluable tools in maintaining a consistently high quality level. Thus, a new dimension in quality and flow control is added to exceptional research, production and inspection facilities; more reasons why the world's leading manufacturer of custom-built and stock delay lines is . . .

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Crystals are grown by a modified Czochralski technique.

Large Diameter
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Cut Domes and Lenses

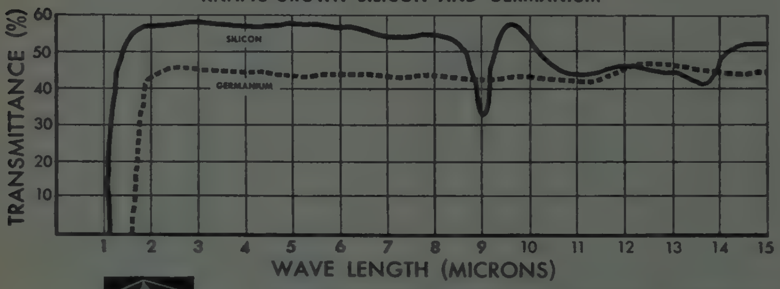
Silicon cut domes and lenses to 8″ diameter, with IR transmission to 97% (coated), are now available in production and evaluation quantities. Diameters up to 19″ will be available in the near future.

These significantly larger diameters can now be grown as a result of recent Knapic research and experimental growing programs. Temperature gradient, pressure, and impurity evaporation controls, as well as unique growing methods, are the result of original Knapic laboratory work.

Germanium lenses and domes are also available

SPECIFICATIONS	SILICON	GERMANIUM
• Hardness	750-2000 (Knoop) Excellent	692 (Knoop) Excellent
• Index of Refraction	3.50 high	4.10 high
• Melting point	1420°c Excellent	958°c Fair
• Density	2.3 gm/cm^3	5.34 gm/cm^3
• Ease of finishing	Excellent—very hard	Good
• Transmission cut-off	About 20 microns Excellent	About 23 microns Excellent
• Reaction to Thermal Shock	Good	Good
• Thermal conductivity	Excellent	Excellent

TYPICAL INFRARED TRANSMISSION CHARACTERISTICS OF UNCOATED KNAPIC-GROWN SILICON AND GERMANIUM



Also manufacturer of Silicon and Germanium crystals for solar cell and semi-conductor devices.

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Western—204 South Beverly Drive, Beverly Hills, California • Phone CR 6-7175
European—#2 Prins Frederick Hendriklaan, Naarden, Holland • Phone K 2959-8988

IRE People

(Continued from page 58A)

Previously, he spent six years with American Telephone and Telegraph's Long Lines Dept. in transmission and equipment engineering positions. In the Navy he served as a communications and electronics officer.

Mr. Culbertson received the B.S.E.E. degree from Purdue University. He is a member of the American Institute of Electrical Engineers and Eta Kappa Nu. He is the author of many papers and articles on telecommunications.

⁂

Dr. Donald A. Dunn (S'46–A'52–SM'56) has been named Director of the newly formed Research Division of Eitel-McCullough, Inc., manufacturer of Eimac electron-power tubes.

D. A. DUNN

The new division, consolidating into an expanded program many Eimac research activities, will headquarter in San Carlos, Calif. Conducting new and advanced technological research in present and new areas of company interest, it will contribute new technical knowledge to the product development programs of Eimac and its subsidiaries, according to Executive Vice President Gould Hunter.

Dr. Dunn, a senior research associate and lecturer in the electrical engineering department of Stanford University, joined Eimac in 1959 as manager of the Supporting Research Group. He holds the B.S. degree from California Institute of Technology, class of 1946. In addition, he received the M.S. and Ph.D. degrees in Electrical Engineering and an LL.B. from Stanford.

He is currently vice-chairman of the San Francisco Section of the Institute of Radio Engineers and is former chairman of the San Francisco Chapter of the Professional Group on Electron Devices.

⁂

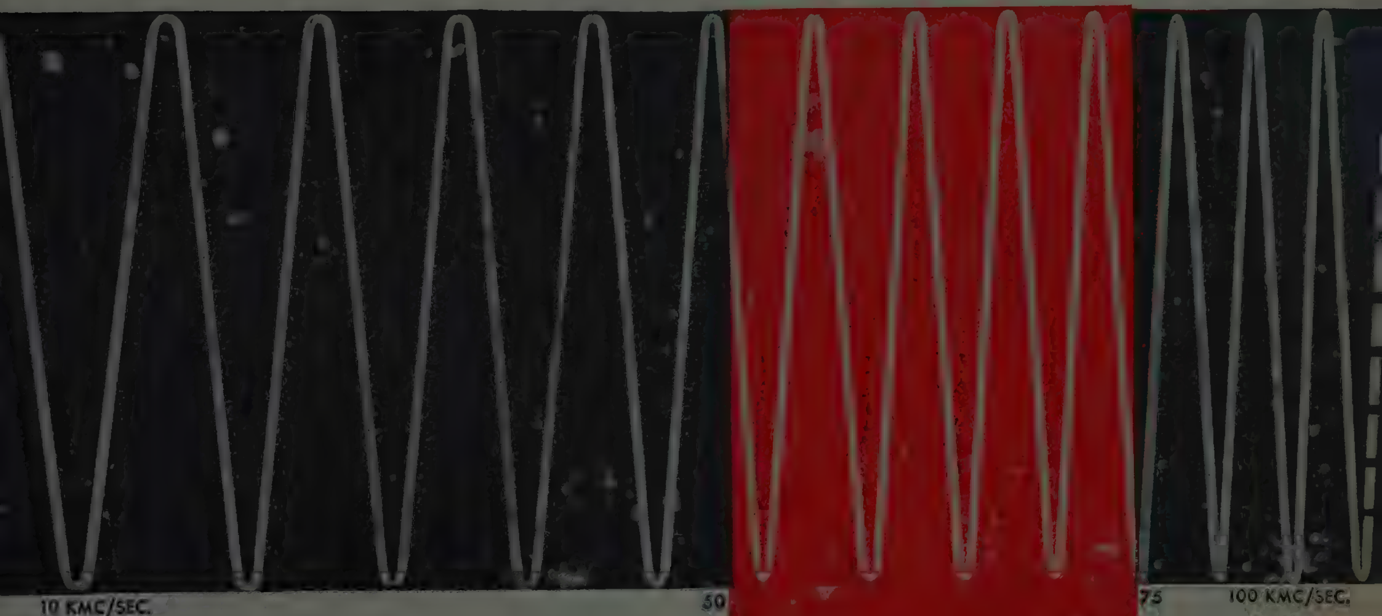
A. P. Fontaine, executive vice president of Bendix Aviation Corporation, has announced the appointments of **Charles M. Edwards** (S'41–A'43–M'45–SM'53) as his assistant and **Dr. G. A. Rosselot** (A'41–M'55) as assistant general manager and associate director of the Research Laboratories division of the company.

Mr. Edwards, formerly assistant director in charge of technical activities of the Research Laboratories division, will be mainly responsible for new product development at the corporate level, in addition to assisting Mr. Fontaine in other corporation activities, including engineering and research, sales, and patent activities.

In 1951 he joined the staff of the Research Laboratories division as a project engineer in charge of the development of

(Continued on page 62A)

Now...PHILCO offers the only commercially available fully-tested mixer diode in the 70 KMC BAND


10 KMC/SEC.
50
75
100 KMC/SEC.

For Long-Range Space Communications and High Resolution Radar Applications

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It is primarily designed for high resolution radar applications and for long-range high altitude or space communications . . . atmospheric absorption prevents jamming from the ground. The Philco 1N2792 is also well suited for EHF video detector applications.

Philco design and application facilities are at your disposal in developing millimeter diodes to meet your specific requirements. For complete information, write Special Components Dept. IR 460.

Test Frequency..........................69,750mc

	TYP.	MAX.	
Noise Ratio, NRo..............	2.0	2.5	times
Conversion Loss, Lc (Note 1).....	8.4	10	db
RF Impedance, VSWR..........	1.35	2.0	times
Crystal noise figure, NF (Note 1)..	11.5	13	db

Note 1: Based on a noise temperature of 5800° K for Roger White Noise Source No. GNW-V18.

MIXER DIODE

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DESIGNED FOR YOUR APPLICATIONS

As represented above, Curtiss-Wright is manufacturing Time Delay Relays for your special applications. Included among the features are:

Instantaneous resetting contacts • Chatter-free operation
Voltage and High Temperature compensation • Snap action
Hermetically sealed • Resistance to severe shock and vibration
Multiple load contacts

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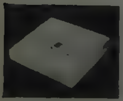
Our application engineers are available to assist you with your individual time delay requirements. Many of the features indicated above are presently included in our standard relay line.

ULTRASONIC DELAY LINES Curtiss-Wright produces Magnetostrictive Delay Lines featuring delays from 5 to 12,000 microseconds, small size, hermetically sealed, and light in weight. Delay lines are designed to meet your application.



COMPONENTS DEPARTMENT • ELECTRONICS DIVISION
CURTISS WRIGHT
CORPORATION • EAST PATERSON, N.J.

TIME DELAY RELAYS • DELAY LINES • ROTARY SOLENOIDS • DIGITAL MOTORS • TIMING DEVICES • DUAL RELAYS • SOLID STATE COMPONENTS

IRE People

(Continued from page 60A)

the analog computer and control equipment for the Bendix three-dimensional flight systems simulator. In 1955 he was named to head the computer department of the division.

He received the B.S. and M.S. degrees in electrical engineering from Massachusetts Institute of Technology, and for five years supervised graduate students at M.I.T. He is a member of the Engineering Society of Detroit.

Dr. Rosselot, formerly assistant director in charge of administration of the Research Laboratories division, joined the staff of the Bendix Products division, missile section, in 1953, as staff assistant and director of engineering. For three years, before being appointed to the staff of the Research Laboratories division, he was director of university and scientific relations for the Corporation.

From 1934 to 1953 he was associated with the Georgia Institute of Technology as director of research. In 1936 he received the Ph.D. degree in physics from Ohio State University. He is a graduate (cum laude) of Otterbein College. He is a member of The American Society for Engineering Education, and the Engineering Society of Detroit.

❖

(Continued on page 64A)

DEPENDABLE FREQUENCY CONTROL

WITH BASSETT CONTROL CRYSTALS for

- CITIZENS BAND USE
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- TAXICAB and POLICE
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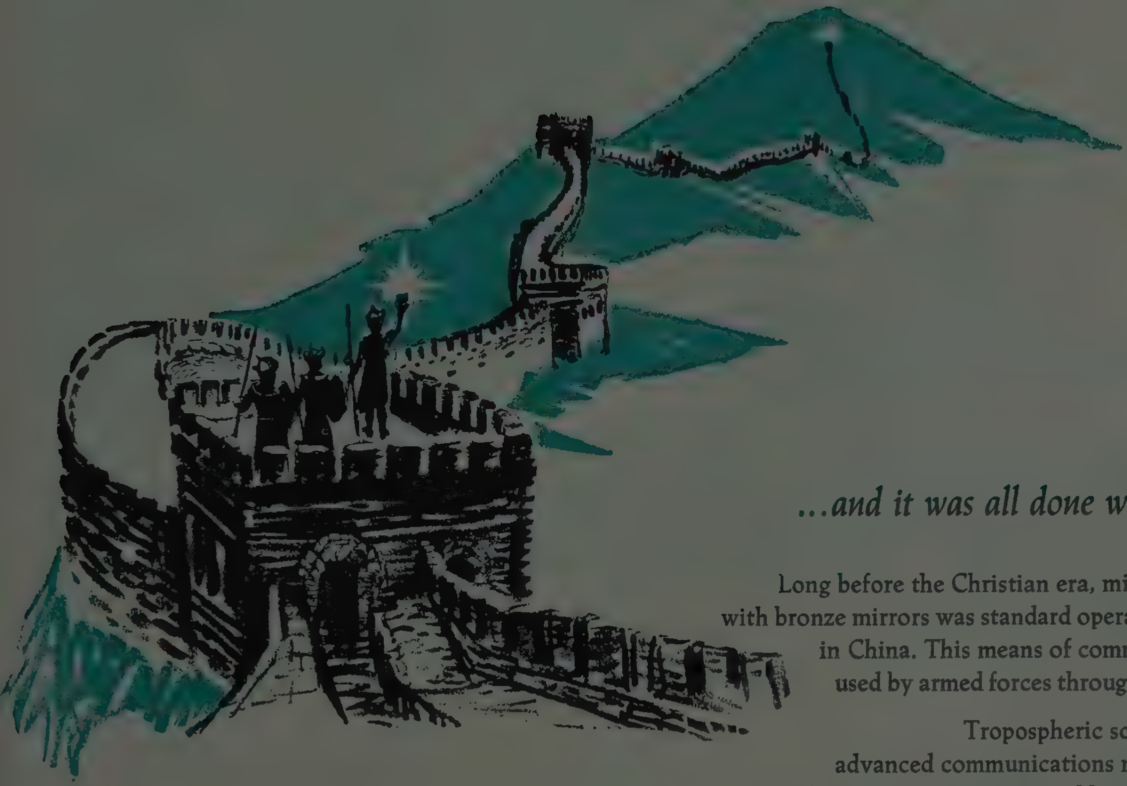

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Long before the Christian era, military signaling with bronze mirrors was standard operating procedure in China. This means of communication was used by armed forces through the centuries.

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The name Radio Engineering Laboratories, too, has been carried afar. REL is a pioneer in the design and construction of tropo scatter communications equipment. You'll find its name on the radio apparatus in use or on order by eight out of nine major tropo networks.

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35 amps
100 amps
150 amps
200 amps
250 amps

Tarzian high-current line combines thermal efficiency with mounting versatility and optional base polarity

The low junction current density of Sarkes Tarzian's high-current silicon power rectifiers results in longer, more reliable operating life. Compare these key Tarzian values with those of other comparably rated units, and you'll see why Tarzian rectifiers have won such wide acceptance among designers:

DC CURRENT	JUNCTION SIZE	THERMAL GRADIENT (Junction to base)	JUNCTION TEMP. RISE
35 amps	.375 inch	9° Centigrade	60°C Maximum
100 amps	.75 inch	5° Centigrade	60°C Maximum
150 amps	.875 inch	7° Centigrade	60°C Maximum
200 amps	1.0 inch	9° Centigrade	60°C Maximum
*250 amps	1.125 inch	11° Centigrade	60°C Maximum

*Available with stud mounting only

In addition to providing for maximum cooling and larger junction area, Tarzian's unique case styling produces a compact, easily mounted rectifier available in flush or stud mounting types. Tarzian high-current silicon power rectifiers are also available from stock in your choice of negative or positive base polarity.

For complete specifications and ordering information, contact your Sarkes Tarzian sales representative or write to Section 4574 F, Sarkes Tarzian, Inc., Semiconductor Division, Bloomington, Indiana.

SARKES TARZIAN, INC.
SEMICONDUCTOR DIVISION
BLOOMINGTON, INDIANA
In Canada: 700 Weston Rd., Toronto 9, Ontario
Export: Ad Auriema, Inc., New York City

IRE People

(Continued from page 62A)

James C. Elms (SM'56) has been named to the position of vice president and general manager (defense operations) for the Crosley Division, Avco Corporation. He will have line responsibility over all activities in the Crosley Division relating to military business, and will coordinate and direct the various activities as they relate to individual military programs. Mr. Elms was vice president of ground electronics and communications for the Division prior to his new appointment.

J. C. Elms

Before joining Avco's Crosley Division ten months ago, he was with the Denver Division, Martin Company, as manager of its Avionics Department where he was responsible for the development of guidance and control, electronic, electrical and ordnance equipment for the Titan ICBM. Before joining Martin, he was manager of the fire control engineering department, Autonetics Division, of North American Aviation, Inc., Downey, Calif.

During World War II, he served as an officer in the United States Army Air Force and was assigned to the Air Service Command and then to the Armament Laboratory at Wright Field, Dayton, Ohio.

Mr. Elms is a native of East Orange, N. J. and Phoenix, Ariz. He holds the B.S. degree in physics from the California Institute of Technology and the M.A. degree in physics from the University of California at Los Angeles. He also served as Research Associate at UCLA in the field of geophysics. He holds memberships in the American Rocket Society (ARS), and the Institute of Aeronautical Sciences (IAS).

❖

The promotion of **William Glass** (A'54) to Manager of the Dayton, Ohio District Office, has been announced by the Ampex Data Products Company, Redwood City, California. A division of Ampex Corporation, the company manufactures magnetic tape recorders widely used for the acquisition, storage and processing of business and scientific data. The equipment plays a major part in aircraft and missile testing. In assuming his new position, he will supervise technical sales and customer relations for the area.

W. Glass

Before joining Ampex, he was with the Applied Science Corporation of Princeton (ASCOP). Earlier he did telemetry systems work for RCA's Patrick Air Force Base, Florida installation, and was a re-

(Continued on page 66A)

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Overall accuracies from ±.05% to ±.01% over −55°C to +85°C range, and to ±.001% from zero °C to +75°C, **without** use of ovens.

Silicon and germanium transistorized. Sinewave, squarewave and pulse outputs. 18, 20, 24, and 28 volt DC inputs.

Conservatively designed **reliable** units, potted in silicone rubber and hermetically sealed, for operation under **MIL** environmental conditions.

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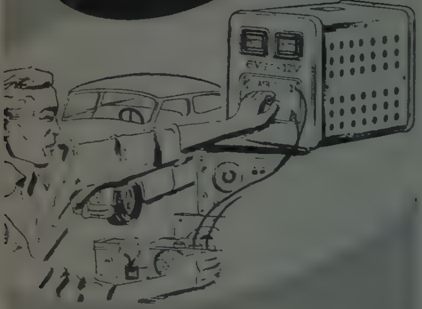
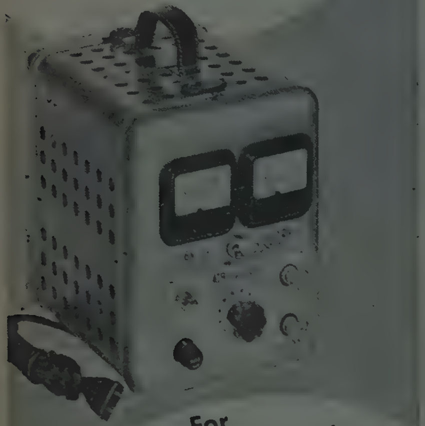


"A" BATTERY ELIMINATORS



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TRANSISTOR OR VIBRATOR OPERATED
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New Models . . . Designed for testing D.C. Electrical Apparatus on Regular A.C. Lines. Equipped with Full-Wave Dry Disc Type — Rectifier, Assuring Noiseless, Interference-Free Operation and Extreme Long Life and Reliability.

TYPE	INPUT A.C. Volts 60 Cycles	D.C. OUTPUT VOLTS	D.C. OUTPUT AMPERES Cont.	D.C. OUTPUT AMPERES Int.	SHIP. WT.	USER PRICE
610C-ELIF	110	6 -or- 12	10 6	20 12	22	$49.95
620C-ELIT	110	6 -or- 12	20 10	40 20	33	$66.95

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✓NEW MODELS ✓NEW DESIGNS ✓NEW LITERATURE
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Quality Products Since 1931
SAINT PAUL 1, MINNESOTA, U. S. A.

IRE People

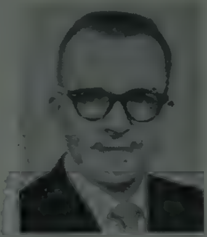
(Continued from page 64A)

search and development engineer for Hughes Aircraft, Culver City, California.

Mr. Glass attended the University of Illinois, and is active in the IRE's Committee on Telemetry and Remote Control.

❖

Dr. Charles K. Hager (M'55) has been named section manager of Automatic Controls for Temco Electronics, a division of Temco Aircraft Corporation.



C. K. HAGER

Dr. Hager received the B.S., M.S. and Ph.D. degrees from the University of Texas. Prior to joining Temco in June, 1959, he was director of applications research at Varo Manufacturing and previously had been employed at Convair and Mandrel Industries. Before his promotion to section manager responsible for design and development of automatic controls and control systems, he was a scientist in Temco's development and analysis group.

❖

General Electric's Communication Products Department here has named **Elmer W. Hassel** (M'49) as manager of Tone-Signalling and Filter Design Engineering.

E. W. HASSEL

The position, newly-created, involves responsibility for design of terminal communication equipment, quartz crystals and filters for General Electric's two-way radio, microwave and carrier current products.

With G-E since 1951, he has specialized in the design and development of selective calling communication devices and tone signalling equipment. He has an extensive background in the field of radio communications, telephone and telegraphy.

Following his graduation from Milwaukee School of Engineering in 1943 with the B.S. degree in electrical engineering, he joined the staff of the Office of the Chief Signal Officer, Wire Communications Section, Engineering and Technical Service, Washington. There, he was engaged in studies related to the development and testing of carrier telephone terminals and repeaters, voice frequency telegraph terminals and repeaters, speech-plus-duplex equipment and amplifiers for the military service.

From 1946 to 1951, he was a faculty member of the Milwaukee School of Engineering where, immediately prior to his association with General Electric Company, he was professor of electrical engi-

(Continued on page 69A)

SEALED IN by HERMETIC HEADERS

Hermetic Seal Corp. Glass-to-Metal and Ceramic-to-Metal Seals for *EVERY* Electronic Use!

- Crystal Bases
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- Single Lead Terminals
- Refrigerant Seals

Hermetic Seal Corporation is your prime source for seals in all sizes, terminations and shapes to fit your specifications. Hermetic is the pioneer in this specialized field and the originator of more than 10,000 different seals. Precision quality-control from design, manufacture and precious metal plating is assured.

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A SENSITIVE RESEARCH

TRIPLE PLAY!

1. AC-DC WIDE RANGE CALIBRATOR

SRIC's versatile AC/DC Volt-Amp-Milliammeter "Self Checking" Model Thach is now offered in combination with a general purpose power supply for use as a convenient, economically priced, wide range instrument calibrator. The Model Thach has 19 overlapping full scale ranges of current and voltage. Operation is almost always at or near full scale for maximum accuracy. The power supply (Model Thach-PS) is a regulated source with its controls interlocked with the Model Thach. Power to the Thach is automatically shut off when ranges are switched, reducing the danger of accidently overloading the calibrator and/or the instrument under test. Output covers the full range of the Thach for DC and 60 cps. 50-2500 cps. can be plugged in for operation over a broader frequency range. Resolution of the power supply controls is better than .05% for any current and voltage range.

Any Model Thach with range combs. 1x or 3x is available for operation with the power supply. In addition, existing instruments in the field can be modernized for this same purpose at a small surcharge.

2. AC-DC LABORATORY STANDARD

The Model Thach can be used independently as a true RMS measuring instrument powered by any appropriate source. It checks its own accuracy against an internal standard cell, and is furnished completely self contained with a high sensitivity galvanometer. General specifications are as follows: Accuracy ± .2% F.S. (.1% at self-check points). Ranges 10/20/50/100/200/500 ma; 1/2/5 amps; 1/2/5/10/20/50/100/200/500/1000 v. Frequency range DC and 7 cps. to 4 kc. Sensitivity 100 ohms volt. Thermocouple D'Arsonval type. 100 division 6.3″ hand drawn mirrored scale. Temperature compensated from 20°C.—30°C. 1000% manually operated overload protection system. Independent of influence of stray magnetic fields. Diamond Pivoted with sapphire shock mounted jewels. Furnished with 1 or 2 thermocouples. Available in two combinations: 1x is self checking at full scale only: 3x is self checking at 5 cardinal points including full scale. Size 16⅜″ x 10½″ x 8¾″ deep. Weight approx. 21 lbs.

Model THACH AND THACH-PS

3. AC-DC GENERAL PURPOSE POWER SUPPLY

The Model Thach-PS, while designed primarily for operation with the Model Thach, can be used independently as a power source for any appropriate application. It only has to be disconnected from the Model Thach to become an excellent general purpose switch controlled laboratory supply.

SPECIFICATIONS:

OUTPUT: DC or AC (60 cps normal; 50-2500 cps depending on input) **Voltage:** Delivers 0-120% at full load and a max. of 0-175% at no load. 1/2/5 v. at 300 ma.; 10/20/50 v. at 100 ma.; 100/200/500/1000 v. at 25 ma. **Current:** 0-5 amps. at 15 watts max. on low power, or 75 watts max. on high power.

INPUT: 90-135 v., 60 cps single phase, or 115 v. 50-2500 cps from a regulated source.

WAVEFORM: 60 cps distortion 2.5% total harmonic content plus line distortion.

RIPPLE: Varies from .5% at full voltage and low currents to a maximum of 2.5% for currents of 1 ampere or less. At currents above 1 ampere, ripple can always be 3% or less if series resistance is sufficient in external circuit and will be no worse than 7% for any possible operation.

RESOLUTION: .05% or better by means of coarse, medium and fine controls.

SIZE: 12¾″ x 17″ x 9″ deep.

WEIGHT: Approx. 59 lbs.

S Symbol of Quality

SENSITIVE RESEARCH INSTRUMENT CORPORATION

NEW ROCHELLE, N. Y.

ELECTRICAL INSTRUMENTS OF PRECISION SINCE 1927

THE
X-15:
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FOR
MAN
IN
SPACE

As North American's X-15 — world's most advanced manned research craft—parts the curtain of earth's atmosphere, the arts of guidance and direction must play a critical role. Sperry's Air Armament Division, assigned the Flight Data System responsibility for the X-15, is meeting the challenge with inertial guidance gear of advanced design, precision and dependability.

But the problems of inertial guidance are not new to Sperry. During the past ten years, over 25-million Sperry man-hours have been employed to develop and produce successful inertial guidance. As a result, the nation has in the Convair B-58 Hustler the most thoroughly studied, analyzed, tested, evaluated and understood inertial guidance system in being — plus the advanced guidance equipment for the X-15 and for other future applications.

And in addition to work on government sponsored space guidance systems and techniques, Sperry scientists and engineers are exploring new and exotic techniques for gyros, advanced miniaturized digital computers, acceleration sensors, zero gravity environment systems—in many cases involving radical departures from current technology—with the aim of developing concepts, systems and hardware that are *ahead* of the challenges of man in space.

SPERRY

AIR ARMAMENT DIVISION, *SPERRY GYROSCOPE COMPANY, DIVISION OF SPERRY RAND CORPORATION, GREAT NECK, NEW YORK*

(Continued from page 66A)

neering and head of the electronics branch of the department of electrical engineering.

Mr. Hassel is a member of the American Institute of Electrical Engineers. He is also affiliated with the Lynchburg Society of Engineering and Science.

❖

Frank J. Hierholzer, Jr. (M'58) has been named Assistant Department Head of the Microtronics Department of Sperry Semiconductor, South Norwalk, Conn.

F. J. Hierholzer, Jr.

He brings to Sperry extensive experience in semiconductors. He was Senior Engineer in the Solid State Electronics Engineering Section of the Materials Engineering Department of Westinghouse Electric Corp. He received the B.S. degree in Electrical Engineering, with honors, at Virginia Polytechnic Institute, the M.S. degree at Stevens Institute of Technology, and is completing the Ph.D. requirements at the University of Pittsburgh.

An associate member of American Institute of Electrical Engineers, Mr. Hierholzer has presented a technical paper to AIEE on "Linear Power Amplifier Using Dynistors or Trinistors." He has several patents in the fields of arc discharge devices and semiconductor applications.

❖

Watkins-Johnson Co. has announced additions to its technical staff engaged in research and development on microwave tubes and devices.

Dr. Boyd P. Israelsen (S'53–A'54–S'58), whose attention will be directed toward the development of low-noise traveling wave tubes, was a National Science Foundation Fellow and research assistant at Stanford Electronics Laboratories. Before coming to Stanford, where he was awarded his Ph.D., Dr. Israelsen was a senior research engineer at Jet Propulsion Laboratory, Pasadena, working on components for missile guidance systems. He has the B.S. and M.S. degrees from California Institute of Technology.

Dr. Kenneth L. Kotzebue (S'56–M'59), who will work on solid-state and electron beam parametric amplifiers, was formerly a senior engineer with the Apparatus Division of Texas Instruments, Inc., Dallas. He has made significant contributions in the field of parametric amplifiers. He was a Bell Telephone Laboratories Fellow at Stanford University, where he received his Ph.D. last year, and a research assistant at Stanford Electronics Laboratories. His B.S. degree is from the University of Texas and his M.S. degree is from the University of California at Los Angeles.

❖

(Continued on page 70A)

NEW

RADIO INTERFERENCE — FIELD INTENSITY MEASURING EQUIPMENT, 375 mc to 1000 mc

The NEW NM-52A RI-FI instrument developed by STODDART to government specifications is now ready for immediate delivery.

Its purpose is to investigate, analyze, monitor and measure to the highest practical degree conducted or radiated electromagnetic energy to military specifications within the frequency range of 375 mc to 1000 mc. In addition, the NM-52A is valuable as a highly sensitive frequency-selective voltmeter and receiver for numerous laboratory and field applications.

OUTSTANDING FEATURES

SENSITIVITY OF 1 MICROVOLT ACROSS 50 OHMS, provides up to 40 db more than Military Measurement Requirements.
SINGLE KNOB TUNING.
RAINPROOF, DUSTPROOF, RUGGEDIZED AND TOTALLY ENCLOSED, for all-weather field use or precise laboratory measurements.
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NEW POWER SUPPLY, 0.5% REGULATION, for filament, bias and plate voltages, and also for use as a standard laboratory power supply.
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TWO DECADE LOGARITHMIC METER SCALE, increases range of voltage measurement without change of attenuator steps.
THREE DETECTOR FUNCTIONS, for peak, quasi-peak or average measurements.
PORTABLE OR RACK MOUNTING, no modification required for laboratory, mobile, airborne or marine installation.
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VISUAL PEAK THRESHOLD INDICATOR, for accurate slide-back peak voltage measurements.
CONSTANT BANDWIDTH OVER ENTIRE FREQUENCY RANGE.

The NM-52A now joins the family of STODDART government approved RI-FI instrumentation covering the frequency range of 30 cps to to 10.7 kmc to provide the finest RI-FI measuring equipment.

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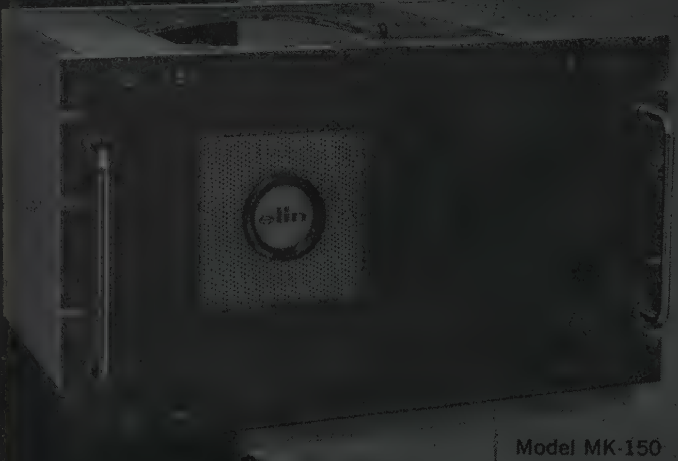

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IRE People

(Continued from page 69A)

Formation of an Antenna and Microwave Component Development Group at Granger Associates here has been announced by Dr. John V. N. Granger, president.

Centralized in the new section will be the firm's activities in the development and manufacture of large parabolic reflector antennas, wide-band log periodic antennas, antennas for missiles and aircraft and associated devices such as multiplexers and band-separation filters.

Dr. Granger said recent additions to the engineering staff in this technical area, together with new development and fabrication facilities recently added, will increase his firm's emphasis on antenna systems and associated components and allow the initiation of new programs in steerable arrays, direction-finding systems and microwave circuitry.

Joining Granger in December to head the new group was **Dr. Ray Justice** (A'55–SM'58), formerly supervisor of research and development on the Radiation Research and Development Section at Convair Division of General Dynamics, San Diego, Calif.

At Convair he directed design and development work in the radomes, antennas, microwaves, solid-state electronics and infra-red and optics groups.

From 1944 until joining Convair in 1956, he was at Ohio State University. There he served at various times as an instructor in the Departments of Mathematics and Electrical Engineering. Primarily he was a research associate at the Antenna Laboratory of the Ohio State University Research Foundation, conducting basic studies in the fields of microwave antennas, arrays and electromagnetic scattering.

Dr. Justice is a member of Sigma Xi and Eta Kappa Nu. He is active in the IRE Professional Groups for Antenna and Propagation and Microwave Theory and Techniques. He is the author of various Ohio State University reports and PGAP papers on near-field measurements.

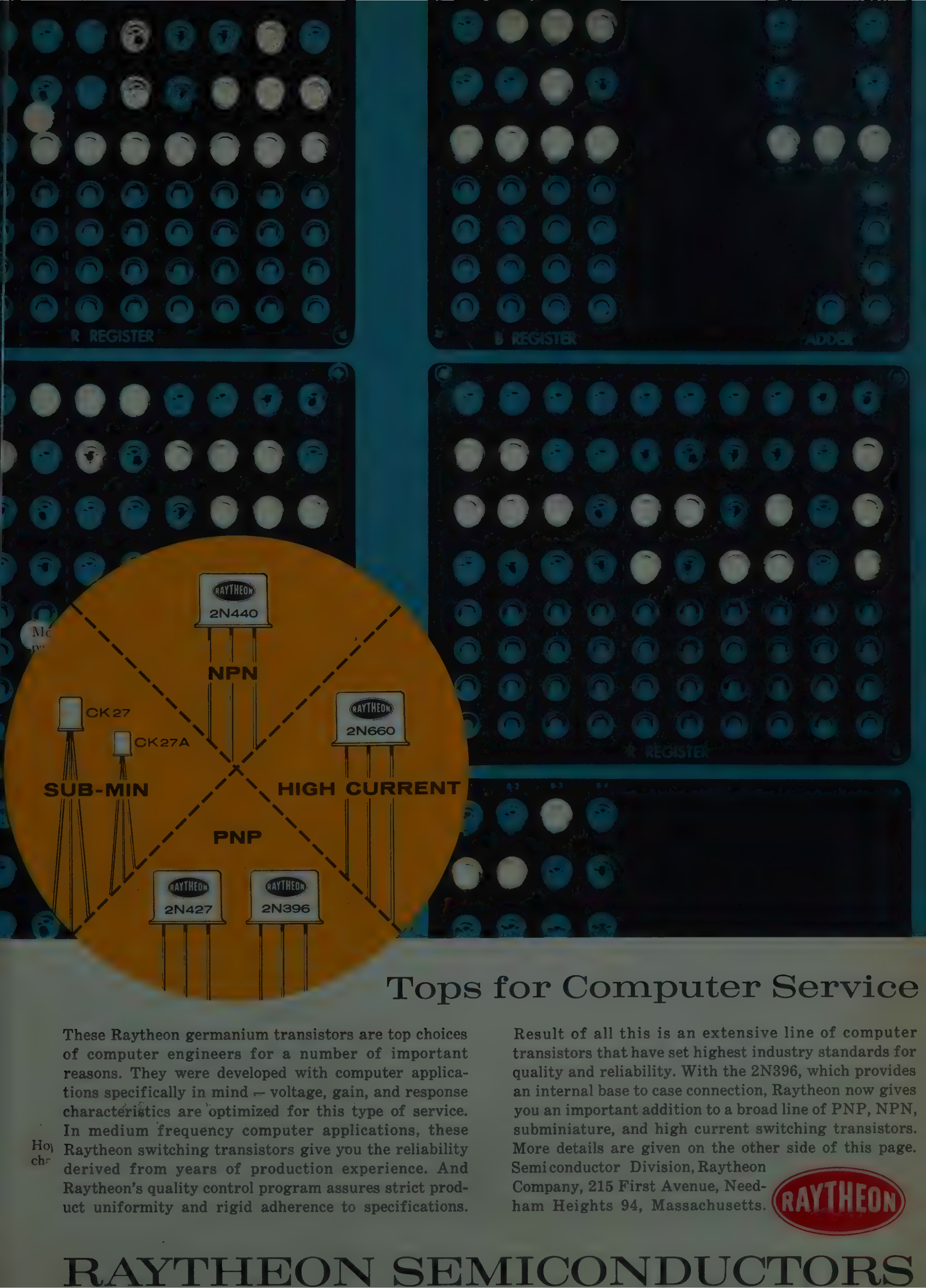
❖

Jack Larsen (M'48–SM'53) has been named Manager of the Special Projects Department by John F. Brinster, President of General Devices, Inc., of Princeton, N. J.

J. Larsen

He is a graduate of Princeton University, the University of Michigan, and George Washington University, with degrees in physics, mathematics, and law. He is a member of the District of Columbia Bar and a registered Patent Attorney.

(Continued on page 73A)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Result of all this is an extensive line of computer transistors that have set highest industry standards for quality and reliability. With the 2N396, which provides an internal base to case connection, Raytheon now gives you an important addition to a broad line of PNP, NPN, subminiature, and high current switching transistors. More details are given on the other side of this page. Semiconductor Division, Raytheon Company, 215 First Avenue, Needham Heights 94, Massachusetts.

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Type	B_{VPT} Max. Volts	$f_{\alpha b}$ mc	H_{FE1}	H_{FE2}	R_{sat} ohms
2N395	15	4.5	40	12	2.2
2N396	20	8.0	60	20	1.3
2N397	15	12.0	80	35	1.1

A COMPLETE HIGH CURRENT SWITCHING LINE PNP germanium switches for 1 amp, high frequency, high gain service. Temperature range −65°C. to + 85°C. Long a production item—excellent availability in large volume.

Type	B_{VPT} Max. Volts	$f_{\alpha b}$ ave. mc	H_{FE1} ave. $I_B = 1mA$ $V_{CB} = -0.25V$	H_{FE2} ave. $I_B = 10mA$ $V_{CE} = -0.35V$	$I_C = 150mA$ ohms
2N658	−16	5	50	40	0.9
2N659	−14	10	70	55	0.6
2N660	−11	15	90	65	0.45
2N661	− 9	20	120	75	0.35
2N662	−11	8	30 min.	50	0.7

IMPROVED DISSIPATION AT LOWER CURRENT VALUES NPN germanium transistors for medium current, high frequency, high gain switching service. Temperature range −65°C. to +85°C. Immediate availability.

Type	B_{VPT} Max. Volts	$f_{\alpha b}$ mc	H_{FE1}	R_{sat} ohms
2N438	25	2.5	25	3.0
2N439	20	5.0	45	3.0
2N440	15	10.0	70	3.0

HIGH RELIABILITY PNP TRANSISTORS germanium transistors for medium current, high frequency switching service. Temperature range −65°C. to +85°C. Immediate availability.

Type	B_{VPT} Max. Volts	$f_{\alpha b}$ mc	H_{FE1}	H_{FE2}	R_{sat} ohms
2N425*	−30	4	30	15	2.2
2N426*	−25	6	40	18	2.2
2N427*	−20	11	55	20	1.3
2N428*	−15	17	80	30	1.1
2N404*	−24	12	See Data Sheet		

* Available to MIL Specification

SUB-MINIATURE TRANSISTORS Sub-miniature transistors for medium current, high frequency, high gain switching. Temperature range −65°C. to +85°C. Immediate availability.

TYPE		B_{VPT} Max. Volts	$f_{\alpha b}$ mc	H_{FE1}	H_{FE2}	R_{sat} ohms
.130″ Dia. x .160″ High	.100″ Dia. x .130″ High					
CK 25	CK 25A	−30	4	30	15	2.2
CK 26	CK 26A	−25	6	40	18	2.2
CK 27	CK 27A	−20	11	55	20	1.2
CK 29	CK 28A	−15	17	80	30	1.1

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IRE People

(Continued from page 70A)

Together with Mr. Brinster, he was one of the organizers of early research in radio telemetering at Princeton University. With the University of Michigan, and later with the Martin Company and the Navy Department, he continued in the general field of missile electronics and control, working on such projects as Matador, Mace, Bomarc, Terrior, Jupiter and Polaris. He served as consultant to the Atomic Energy Commission in the development of electronic portions of reactor controls for the submarines, Nautilus and Sea Wolf. He headed the Navy Department's Missile Guidance Section on the Polaris Project and was honored with the Meritorious Civilian Service Award for his contributions to the Navy Department. Immediately before coming to General Devices, he was a member of the Patent Department of the Bell Telephone Labs.

A member of the New Jersey Patent Law Association and the Acoustical and Physical Societies, Mr. Larsen holds a number of patents in the electronics field.

◆

The appointment of **Paul J. Larsen** (A'37–M'41–SM'43) as director of government relations for Allen B. Du Mont Laboratories, Inc. has been announced by Dr. Allen B. Du Mont, the company's chairman. Mr. Larsen will maintain liaison for Du Mont Laboratories with both civilian and military branches of the government for its military electronic systems and for its commercial electronic products. He will make his offices at Du Mont's Washington, D. C. headquarters, in the Pennsylvania building.

P. J. Larsen

Formerly director of Civil Mobilization and Defense, Executive Office of the President of the United States, he is a veteran of more than 30 years experience in the electronics industry. He comes to Du Mont from Hughes Aircraft Company, where he was manager of East Coast Operations and assistant secretary stationed in Washington. Prior business and professional background include his association with Borg-Warner Corp. as assistant to the president responsible for all governmental defense activities. Before that he was associate director of Los Alamos Scientific Laboratory, and next the director of Sandia Laboratory, and consultant to the President on activities of the newly formed Sandia Corporation.

He was also associated with Johns Hopkins University from 1941–47 and had charge of engineering and production throughout the United States on radio proximity fuses. Prior associations were with Baird Television Corp. of England,

(Continued on page 74A)

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LINEARLY
from 1 mv to 1000 v

Accuracy of conversion
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for frequencies 50 cps to 10 KC

Price $450

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Model 710

The Model 710 Linear AC to DC Converter converts an AC voltage to a DC voltage which can be measured with an accurate DC device such as Type K Potentiometer, Digital DC Voltmeter, Recorder, etc. With such a combination, a wide range of voltages can be measured with up to 0.25% accuracy, which is considerably better than accuracies of present-day vacuum tube voltmeters. Such a system is more sensitive, covers a wider frequency and voltage range, and is much more rugged and foolproof than a laboratory standard instrument of comparable accuracy. It is also adaptable for use by untrained personnel and on production lines.

The instrument covers an input voltage range of 1 mv to 1000 volts which is divided into six decade ranges. For every decade range the DC output varies from 0.1 volt to 1 volt. The input impedance of the converter has a resistive component of 2 megohms shunted by 15 pf to 25 pf, depending on the range.

The output of the Model 710 Converter is a linear function of the input voltage within each decade. A small error may exist in the decading of the input attenuator or in the frequency response of the amplifier. This error does not exceed ±0.25% over a frequency range of 50 cps to 10 KC and ±0.5% over a range of 30 cps to 50 KC. The upper frequency limit of the instrument is 250 KC, at which point the accuracy is ±1%.

The DC output of the converter is single ended and has a maximum output emf of 1 volt with a source impedance of approximately 10,000 ohms. The instrument is the average responding type for distortions as much as 30%, but is calibrated in RMS of a sinewave.

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IRE People

(Continued from page 73A)

Warner Bros. Pictures, and Marconi Wireless Telegraph Co.

He is a recipient of the Presidential Certificate of Merit and Citation and has served on the Technical Advisory Panel on Aeronautics and the Panel on Nucleonics.

A fellow of the Society of Motion Picture and Television Engineers, Mr. Larsen is a member of numerous other professional organizations connected with aviation, missiles, ordnance, and engineering.

❖

A. V. J. Martin

Albert V. J. Martin (SM'56) has left the United States to return to his native country, France. For 3 years he was assistant Professor of Electrical Engineering at Carnegie Institute of Technology, Pittsburgh, Pa.

Dr. Martin has set up a publishing company at 61, rue de Maubeuge, Paris, 9, France. Its first product is a technical magazine entitled "Electronique et Automatisme," the first issue of which appeared in January, 1960.

Dr. Martin is a member of the American Institute of Electrical Engineers, the British Institution of Radio Engineers, Sigma Xi, and Société Française des Radioélectriciens.

❖

M. Nelles

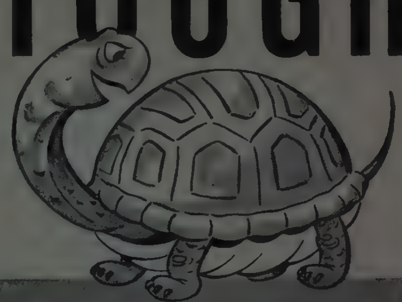
Dr. Maurice Nelles (M'59) has joined American Electronics, Inc., Los Angeles, Calif., to fill the newly-created post of Vice President Engineering, according to Phillip W. Zonne, President of the firm. In this new position, Dr. Nelles will direct the engineering planning for the corporation and its nine divisions on both this coast and the Atlantic seaboard.

Immediately prior to joining American Electronics, Inc., he was Vice President Research and Development and Chairman of the Corporate Product Planning Committee of Crane Company, Chicago. He was also a Director of the Hydro-Aire Company division of Crane, as well as a Director of Crane, Ltd., which operated seven plants in Canada and one in England.

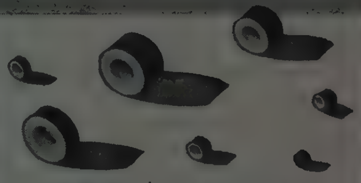
Before his association with the Crane Company, he was Director of Research for Borg-Warner Corporation, and Director of diversification and research development for Technicolor Corporation. He has also served as Professor and Head of the Aeronautical Engineering Department, University of Southern California, Manager of

(Continued on page 76A)

TOUGH



...AS A
TURTLE'S BACK



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(Continued from page 74A)

Research, Allan Hancock Foundation, and Professor and Director of the Engineering Experiment Station at Pennsylvania State University.

During World War II, he was Deputy Director of the War Production Board, and Chief Engineer, Office of Production Research and Development. He is a member of American Society of Mechanical Engineers, American Society of Tool Engineers, Institute of Aeronautical Sciences, Society of Automotive Engineers, American Rocket Society, America Society of Metals, American Institute of Naval Engineers, American Institute of Mining, Metallurgical and Petroleum Engineers, and Trustee of the Midwest Research Institute.

Dr. Nelles received the B.A. and M.A. degrees from the University of South Dakota, and a Ph.D. degree from Harvard University.

⁂

G. Emerson Pray (M'41–SM'53) has been elected a vice president of Ling-Altec Electronics, Inc. to direct the company's activities related to the national defense program, it was announced today by James J. Ling, chairman of the board.

G. E. Pray

He will operate from the Washington, D. C. office to facilitate the company's negotiations of prime and subcontracts for its 14 operating subsidiaries and divisions engaged in governmental work.

A licensed professional engineer and physicist with a background of scientific work and experience in company management, Mr. Pray recently resigned as assistant to the president and director of government operations of Dresser Industries, Inc.

He holds an electrical engineering degree from Rensselaer Polytechnic Institute and the M.S. degree in physics from Rutgers University. He has patented many of his inventions in the radio, electronic and sonic fields, and is a well-known authority and lecturer in these areas of scientific research and development.

⁂

Dr. James A. Van Allen (SM'59–F'60), discoverer of the radiation belts surrounding the earth which bear his name, has been named by the Institute of the Aeronautical Sciences to receive the second annual Louis W. Hill Space Transportation Award.

The award was established by the Louis W. and Maud Hill Family Foundation of St. Paul for the purpose of "encouraging and supporting fundamental research and the early application of new knowledge in behalf of the welfare of mankind." One of the top awards in the aerospace field, it includes an honorarium of $5,000 ($10,000 if more than one recipient is named).

(Continued on page 78A)

NOTABLE DEVELOPMENTS AT JPL

HYPERSONIC WIND TUNNEL

One of the latest research facilities at the Jet Propulsion Laboratory is the recently completed, continuous-flow hypersonic wind tunnel. Developed by the Lab, this new tunnel generates air speeds up to ten times the speed of sound. Its 21-inch square test section provides accommodation for models up to four feet long thus permitting increased model instrumentation. This large test section at Mach 10 with a continuous uniform air flow broadens JPL capabilities in the important area of fluid dynamic research.

To minimize structural deflections due to temperature changes and thus the time required to reach equilibrium conditions, the entire tunnel structure is water cooled and housed in an air conditioned building. Any Mach number between five and ten can be precisely set by means of flexible stainless steel nozzle plates that are positioned to a ten-thousandth of an inch. Calibration results indicated satisfactory solution of the design problems encountered.

The high speed data-acquisition, reduction, and presentation system is designed for high production testing of the nation's most advanced missiles and re-entry configurations. Stability and control phenomena in new regimes can be studied experimentally under carefully controlled conditions.

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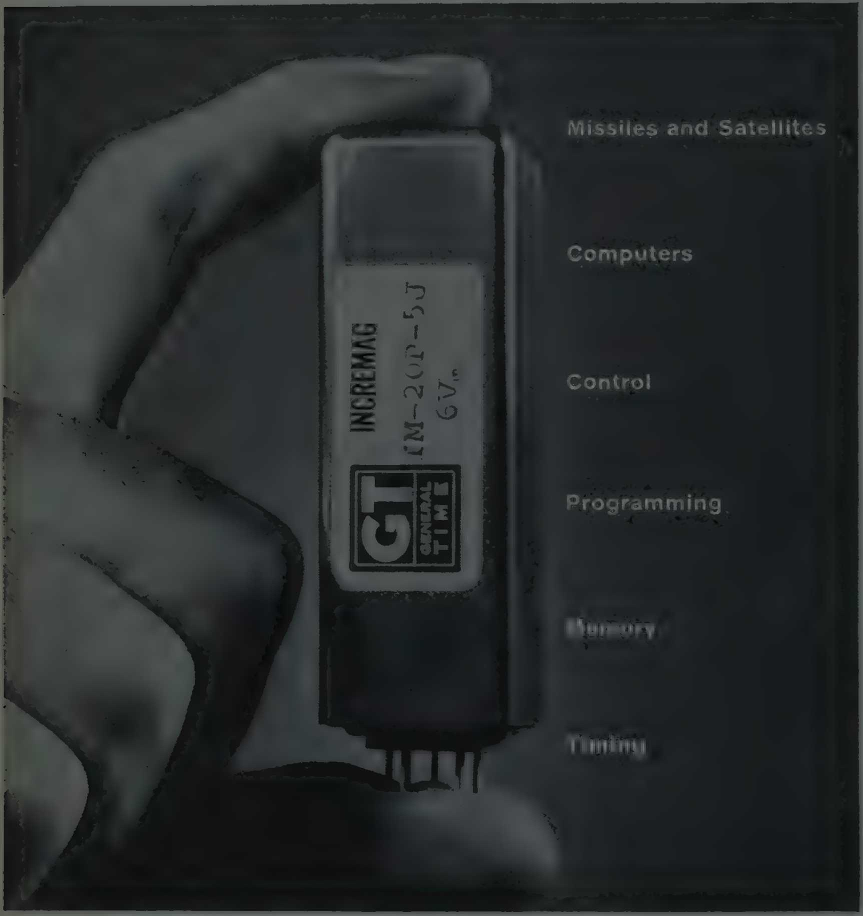

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IRE People

(Continued from page 76A)

Dr. Van Allen, who heads the department of Physics and Astronomy at the State University of Iowa, detected the radiation belts on the basis of data obtained by Explorer I. The award citation states "By combining simple and direct techniques with great ingenuity, he established beyond doubt the existence, intensity and extent of the radiation belts above the earth's surface that now carry his name."

Dr. Van Allen will receive the Hill Award at the annual Honors Night Dinner of the Institute of the Aeronautical Sciences, which will be held at the Astor Hotel in New York on January 26, 1960. He will be present in person to accept.

❖

Varian Associates has announced the appointment of **Willis H. Yocom** (A'54–M'57) as Manager, Wave Tube Development. He will head research and development of all wave tubes.

W. H. YOCOM

He received the B.A. degree in physics from Oberlin College, the B.S. degree in electrical engineering from M.I.T., and the M.S. degree in electrical engineering from Stevens Institute of Technology. From 1942 to 1956 he did research and development work for Bell Telephone Laboratories. He joined Varian Associates' Tube Division in 1956.

Mr. Yocom holds five patents in the microwave tube field including the Basic patents on slalom focusing. He is a member of Sigma Xi.

❖

Homer A. Ray, Jr. (A'45–M'50–SM'55) has been selected for the newly-created position of Engineering Assistant at Rixon Elctronics Inc., according to C. J. Harrison, Vice President and Operations Director. In this capacity, he will assist Mr. Harrison in the engineering phases of planning and production operations.

H. A. RAY, JR.

Prior to joining ing Rixon, he was Manager of the Engineering Division, Smith Electronics, Inc. There, he assisted in the development, engineering, drawings, and specifications for the Voice of America's proposed new East Coast Facility. Before that he was Chief Engineer and Director of Photogrammetry, Inc., making optical and electronic instruments primarily for the Navy.

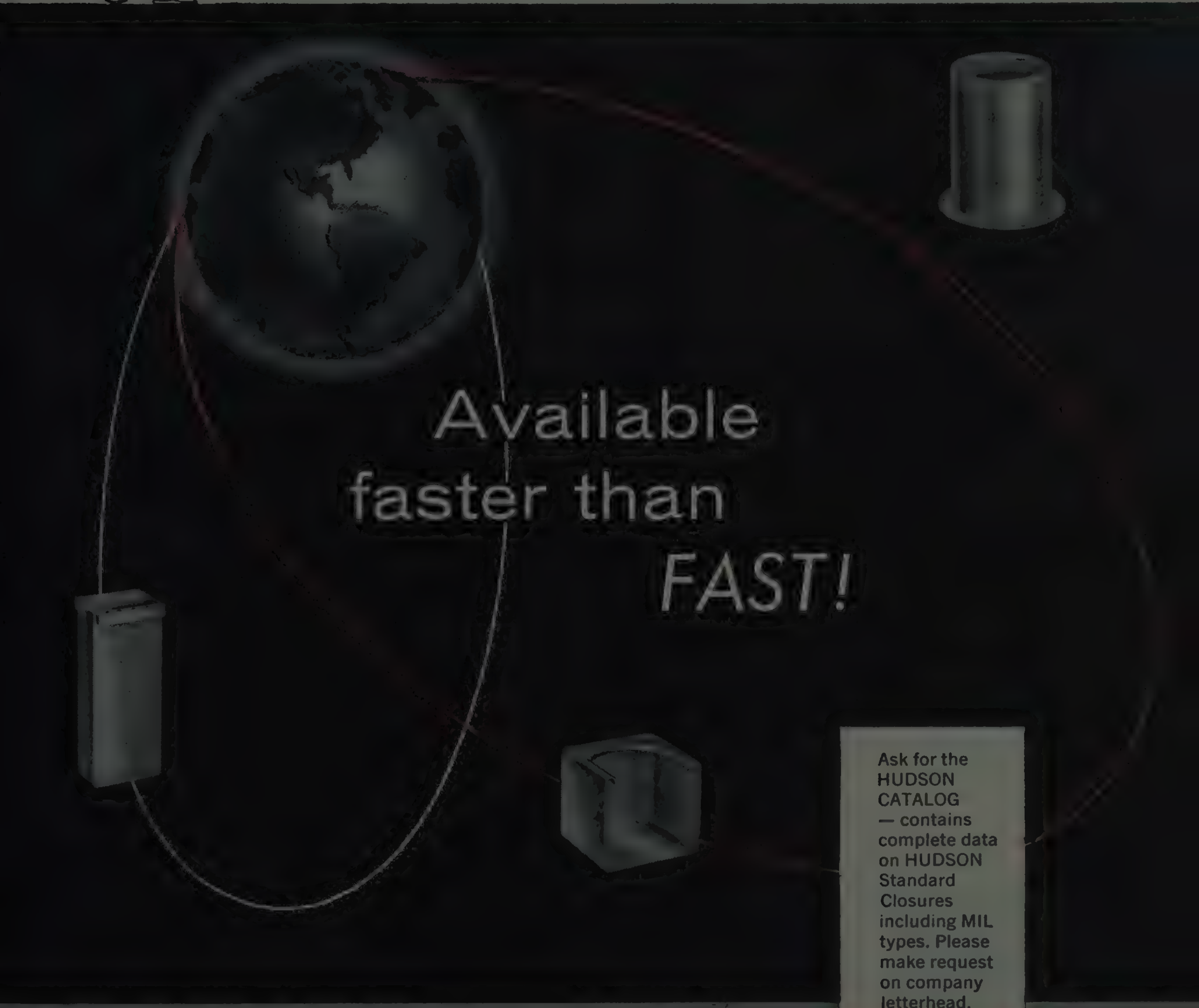
With Page Communications Engineers from 1955 to 1956, he was Head of the Antenna Section and Assistant Manager of

(Continued on page 80A)

Vital to your design orbit!

HUDSON PRECISION-DRAWN CLOSURES

Precision Standard Tooling makes the Difference – The HUDSON line provides complete versatility for the designer/engineer. The industry's widest range of round, square and rectangular closures supplied with dozens of modifications to meet unusual applications. Standardized designs assure fast delivery and precision quality – every time! All finishes available on components of mu metal, nickel-silver, aluminum, brass, copper, stainless steel and steel. Call or write for quotations.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Ask for the HUDSON CATALOG — contains complete data on HUDSON Standard Closures including MIL types. Please make request on company letterhead.

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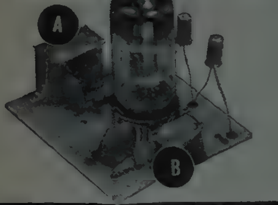


Thin Versatile Co-Netic and Netic Magnetic Shielding Foils

Permit positioning foil-wrapped components A & B closely, minimizing interaction due to magnetic fields . . . making possible compact and less costly systems.

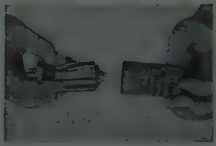
How thin Co-Netic and Netic foils lower your magnetic shielding costs:
1) Weight reduction. Less shielding material is used because foils (a) are only .004" thick and (b) cut and contour easily.
2) Odd shaped and hard-to-get-at components are readily shielded, saving valuable time, minimizing tooling costs.

These foils are non-shock sensitive, non-retentive, require no periodic annealing. When grounded, they effectively shield electrostatic and magnetic fields over a wide range of intensities. Both foils available from stock in any desired length in various widths.

Co-Netic and Netic foils are successfully solving many types of electronic circuitry magnetic shielding problems for commercial, military and laboratory applications. These foils can be your short cut in solving magnetic problems.

Cuts readily to any shape with ordinary scissors.

Wraps easily.

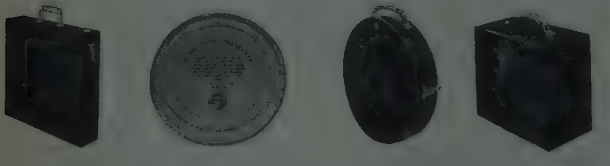

Inserts readily to convert existing non-shielding enclosures.

Shielding cables reduces magnetic radiation or pickup.

Wrapping tubes prevents outside magnetic interference.

PROTECT VITAL MAGNETIC TAPES

When accidentally exposed to unpredictable magnetic fields, presto!—your valuable data is combined with confusing signals or even erased.


For complete, distortion-free protection of valuable magnetic tapes during transportation or storage. Single or multiple reel Rigid Netic Enclosures available in many convenient sizes and shapes.

Thin pliable foil wraps easily around magnetic tape, maintaining original recorded fidelity.

KEEP DOOR CLOSED
Rigid Netic (.014" and up in thickness) Shielded Rooms and Enclosures for safe, distortion-free storage of large quantities of recorded magnetic tapes.

Composite photo to demonstrating that magnetic shielding qualities of Rigid Netic Alloy Material are not significantly affected by vibration, shock (including dropping or bumping) etc. Netic is non-retentive, requires no periodic annealing.

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ORIGINATORS OF PERMANENTLY EFFECTIVE NETIC CO-NETIC MAGNETIC SHIELDING

IRE People

(Continued from page 78A)

the Page Communications phases of the DEW Line Project. He was manager of the engineering and procurement phases of the NATO Project that installed NATO's communications network from Paris through Naples to Izmir, Turkey.

His earlier job experience was with The Queen City Broadcasting Corp. of Seattle, Wash. as Chief Engineer and also as Chief Engineer of the Field Division of Developmental Engineering Corp. doing research and development work on antennas and instruments. Previously, he had a Consulting Engineering Office in Washington, D. C. in partnership with George E. Gautney & Ray, with a practice in broadcast and television engineering.

Mr. Ray was educated at Kent State, Kent, Ohio and the University of Cincinnati and is a Registered Professional Engineer in the District of Columbia. He is associated with Broadcasting Magazine, and is a member of the American Society of Photogrammetry and holds patents in this field.

◆

H. H. SCOTT

Hermon H. Scott (M'35—SM'43—F'52) president of H. H. Scott, Inc., Maynard, Mass., has been elected chairman of the board of directors of the Institute of High Fidelity Manufacturers.

The Institute, an association of leading producers of component high fidelity equipment, was formed in 1955 to educate the public about component high fidelity and to set industry standards. Started with 23 members, the Institute membership now numbers more than 120 leading manufacturers, recording companies and publishers throughout the country.

An electrical engineer with B.S. and M.S. degrees from the Massachusetts Institute of Technology, he is a fellow of the Audio Engineering Society and the Acoustical Society of America. He is credited with the design of the first modern-type sound level meters and the first high quality broadcast station monitoring equipment.

Mr. Scott has published numerous technical papers on the measurement and reproduction of sound and related laboratory measuring equipment and holds patents on many developments in these fields. In 1951 he was cited by the Audio Engineering Society for "outstanding achievement in the field of audio engineering."

◆

Smyth Research Associates, Inc., has named **Louis G. Trolese** (A'30–SM'47–F'60) Vice President for Engineering. This new position reflects the firm's expanded activity in the development of electronic components and systems.

(Continued on page 82A)

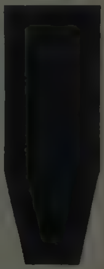
A TALENT FOR BINARY DATA LINK SYSTEMS

The application of binary data link transmission to large or small radio networks is made feasible by the Stromberg-Carlson Automatic Binary Data Link.

The broad capabilities of the equipment include:

- 5-character alphabetical address, providing over 11 million private addresses;
- automatic response to interrogation, which provides a 5-character alphanumeric message useful for indicating any element of a status report such as altitude, heading, position.
- Tape readout in the airborne unit.

Detailed information on request.



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PROVEN RELIABILITY—
SOLID-STATE POWER INVERTERS
over 260,000 logged hours— voltage-regulated, frequency-controlled, for missile, telemeter, ground-support, 135°C all-silicon units available now—

INTERELECTRONICS VOLTAGE REGULATED INTERVERTER POWER SUPPLY

Interelectronics all-silicon thyratron-like gating elements and cubic-grain toroidal magnetic components convert DC to any desired number of AC or DC outputs from 1 to 10,000 watts.

Ultra-reliable in operation (over 260,000 logged hours), no moving parts, unharmed by shorting output or reversing input polarity. Wide input range (18 to 32 volts DC), high conversion efficiency (to 92%, including voltage regulation by Interelectronics patented reflex high-efficiency magnetic amplifier circuitry).

Light weight (to 6 watts/oz.), compact (to 8 watts/cu. in.), low ripple (to 0.01 mv. p-p), excellent voltage regulation (to 0.1%), precise frequency control (to 0.2% with Interelectronics extreme environment magnetostrictive standards or to 0.0001% with fork or piezoelectric standards).

Complies with MIL specs. for shock (100G 11 mlsc.), acceleration (100G 15 min.), vibration (100G 5 to 5,000 cps.), temperature (to 150 degrees C), RF noise (I-26600).

AC single and polyphase units supply sine waveform output (to 2% harmonics), will deliver up to ten times rated line current into a short circuit or actuate MIL type magnetic circuit breakers or fuses, will start gyros and motors with starting current surges up to ten times normal operating line current.

Now in use in major missiles, powering telemeter transmitters, radar beacons, electronic equipment. Single and polyphase units now power airborne and marine missile gyros, synchros, servos, magnetic amplifiers.

Interelectronics—first and most experienced in the solid-state power supply field produces its own all-silicon solid-state gating elements, all high flux density magnetic components, high temperature ultra-reliable film capacitors and components, has complete facilities and know how—has designed and delivered more working KVA than any other firm!

For complete engineering data, write Interelectronics today, or call LUdlow 4-6200 in New York.

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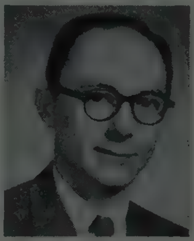
Washington, D. C. Oliver 5-6400

IRE People

(Continued from page 80A)

Peter G. Smee (M'47–SM'52) has been appointed manager of Microwave Systems Design for General Electric's Communication Products Department in Lynchburg, Va. The appointment was announced by I. T. Corbell, the department's manager of Microwave Engineering.


P. G. Smee

Mr. Smee is a native of Lincoln, England and received his technical training in the United Kingdom. He holds several U. S. patents in the communication field. He joined General Electric at Electronics Park, Syracuse, N. Y., in 1946 and has specialized in microwave engineering for the past several years.

❖

John M. Thompson (J'27–A'28–M'47–SM'53) has been named as Vice President and General Manager of ITEMLAB, Inc. located at Port Washington, N. Y. He will head the activities of this concern in the fields of environmental testing, design, research and development.

J. M. Thompson

He has been laboratory chief of the Rome Air Development Center's Test Facility Laboratory for the past ten years and prior to that was associated with Watson Laboratories, Signal Corps Engineering Laboratories and the USVA for a total of 25 years government service.

Mr. Thompson is a senior member of the Instrument Society of America, a member of the American Institute of Electrical Engineers, American Society of Metals and Institute of Environmental Sciences and associate member of the American Society of Heating, Refrigeration and Air Conditioning Engineers.

❖

Use Your
IRE DIRECTORY!
It's Valuable

TARGET... better than 99.99% switching transistor survival

CBS NPN germanium switching transistors are being delivered to a military project specification with an objective of 99.99% survival in 1000 hours. This means that only one transistor in 10,000 can fail due to parameter changes . . . and catastrophic failure rate is even lower.

CHECK THESE SPECIAL RELIABILITY CONTROLS

100% PRODUCTION

- Every crystal identified throughout production
- Die thickness maintained within ± 0.00005 inch
- Contamination eliminated by "dry-box" assembly
- Microscopic inspection of all critical operations
- Aging at 100°C for 200 hours
- Temperature cycling from −65° to 85°C
- Detergent bomb testing at 100 psig
- Radiflo testing to 10^{-12} standard cc/sec

RELIABILITY

- Voltage-biased, 55°C on 1000 hour life
- Storage, 85°C on 1000 hour life
- Intermittent operation, 150 mw for 1000 hour life
- Samplings 100 times larger than MIL-T-19500A
- Product identified with crystal for field survival analysis

ENVIRONMENTAL TESTING

- Shock, 500 G for 1 millisecond
- Vibration, 20 G for 20-2000 cps
- Soldering of leads, 260°C for 10 sec
- Fatigue of leads, 16 oz. pull, three 90° arcs

YOUR INQUIRIES INVITED

Inquiries regarding *your* special reliability requirements for NPN germanium switching transistors are invited. Please call direct . . . or your local sales office.

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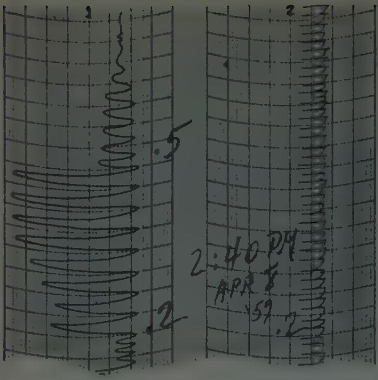

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Time...

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The new Model WWVT receiver, designed especially for remote operations under extreme environmental conditions, is a highly-sensitive crystal-controlled instrument capable of utilizing WWV and WWVH transmission.

Model WWVT

A 6-position dial switches *instantly* to any Standard Frequency—2.5, 5, 10, 15, 20 or 25 mc. It is small, light-weight and rugged—sealed metal case and potted components, all transistorized and battery operated, and has better than 2 mv sensitivity. Priced at $690.00. Model WWVTR (rack mount) $725.00

Send for bulletin #159A which details many free services available from WWV & WWVH.

SPECIFIC PRODUCTS
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Section Meetings

AKRON

"Computer Activities in Europe and Russia," Dr. N. R. Scott, University of Michigan. 1/19/60.

ALAMOGORDO-HOLLOMAN

"Teaching Machines and Tomorrow's Education," Dr. R. F. Mager, Fort Bliss Human Research Unit. 1/18/60.

ALBUQUERQUE-LOS ALAMOS

Conducted tour of Lovelace Radiation Laboratory plus four papers by Dr. White of Lovelace Foundation. 11/17/59.

"For Better Conversations," Dr. J. P. Molnar, Sandia Corp. Fellow Awards Dinner meeting. 2/1/60.

ATLANTA

"Quantum Electronics," Dr. M. W. Long, Georgia Tech. 1/22/60.

BALTIMORE

"Military and Industrial Applications of Infrared," A. H. Canada, Institute for Defense Analyses, Washington, D. C. 2/1/60.

BAY OF QUINTE

"Theory of Relativity," Dr. G. A. Harrower, Queens University. 1/20/60.

CENTRAL FLORIDA

"Infrared Technology & Physics," Dr. S. S. Ballard, University of Florida. 1/14/60.

CENTRAL PENNSYLVANIA

"Fission, Fusion and the Future," F. A. Kramer, Public Service Electric, Gas Co. 11/17/59.

"The USSR Forges Ahead," Dr. J. L. McLucas, Singer, Inc. 12/15/59.

"Trends in Electronic Instrument Design," Roy Mershon, Weston Instruments Co. 1/19/60.

CHICAGO

"Hot Electrons from Cold Cathodes," Dr. A. M. Skellet, Tung-Sol Electric, Inc. 9/11/59.

"SAGE—A Data Processing System for Air Defense," S. J. Hansen and C. A. Zraket, Mitre Corporation. 10/9/59.

"Molecular Engineering, the Future of Electronics," Dr. G. C. Sziklai, Westinghouse Electric Corp. 11/13/59.

"The Place of Ionospheric Scatter in World Communications," W. H. Collins and K. E. Oittinen, Page Communications Engineers, Inc. 12/11/59.

"Circuit Design Using Tunnel Diode Devices," Dr. K. K. N. Chang, RCA Labs. 1/8/60.

CHINA LAKE

"Parametric Amplifiers," Dr. C. J. Carter, Space Technology Lab. 11/9/59.

"Electron Microscopes," Dr. Ernst Bauer, U. S. Naval Ordnance Test Station. 1/11/60.

CLEVELAND

"Thermo Electricity," C. J. Frank, Westinghouse Electric Corp. 1/21/60.

COLUMBUS

Inspection Tour Electronics Laboratories Facility, North American Aviation. Comments by David Leonard of the Company. 2/9/60.

DALLAS

"Tactics of Air Launching a Ballistic Missile for Satellite Reconnaissance," Eaton Adams, Convair Fort Worth. 1/19/60.

"Esaki and Other Special Diodes," G. C. Dacey, Bell Telephone Labs. 2/4/60.

"Information Theory," W. A. Youngblood, University of Texas. 2/11/60.

(Continued on page 88A)

MINCOM

announces a new double-duty magnetic tape system

the CM-100

records & reproduces 6 SPEEDS

Greater bandwidths at lower speeds in both analog and pulse recording/reproducing (for example, 500 kc at 60 ips). All in one compact standard rack. Details? Write for brochure.

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3M COMPANY
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2049 South Barrington Avenue, Los Angeles 25, Calif.

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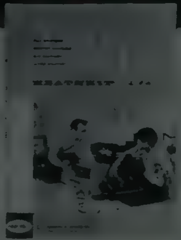
HEATHKIT **1960 CATALOG**



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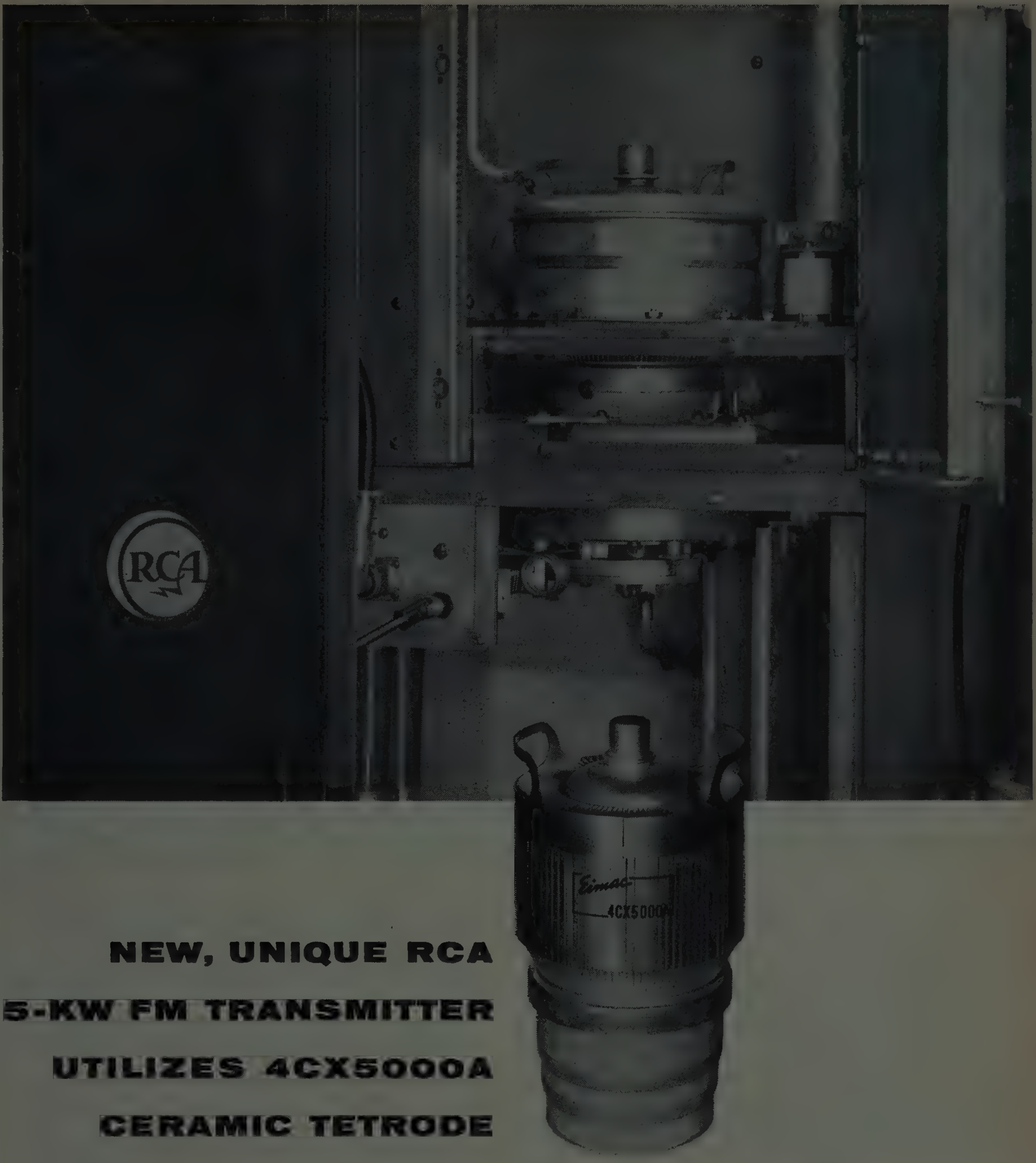


NEW, UNIQUE RCA 5-KW FM TRANSMITTER UTILIZES 4CX5000A CERAMIC TETRODE

RCA has recently developed a unique new 5-KW FM transmitter which utilizes the new technique of multiplexing. This provides simultaneous transmission of two or more program channels on the same RF carrier to meet increased demands of FM stations for additional program services.

The PA stage of the new BTF-5B transmitter is composed of a single Eimac 4CX5000A ceramic tetrode, which produces the 5000-watt output. This tetrode offers high power gain and excellent stability to assure faithful transmission of the broadband multiplex signals.

That's why the 4CX5000A was the logical choice of discriminating RCA engineers. Its many exclusive ceramic design features help to make possible this conservatively rated, high power, air-cooled transmitter.

These ceramic extras are now available in more than forty Eimac tube types—used in many types of communication, pulse and industrial equipment.

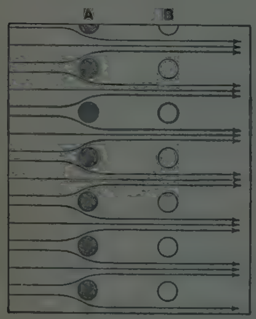
EITEL-McCULLOUGH, INC.

San Carlos, California

Nothing is NEWER than like G-E Shadow Grid... anode...New products New engineering: direct-

MEANS LOWEST-NOISE PENTODE!

The new Shadow Grid tube is an advanced concept applied by General Electric. It makes possible high-gain pentode performance at a low noise level found up to now only in triodes. Electron flow is channeled *between* the wires of the screen grid. There is minimum contact of electrons with grid. Consequently, noise-producing screen current is held to a minimum. A plate-to-screen current ratio of 25 to 1 can be obtained with new General Electric Type 6FG5 for TV tuners.



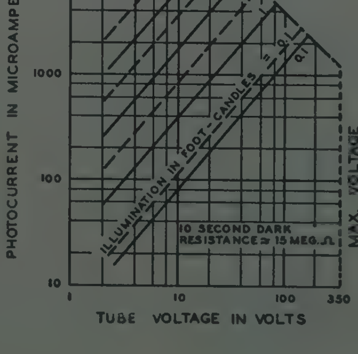
Electron flow from cathode past control grid is guided by electrostatic field in the vicinity of

Shielding grid (A) into streams passing between the wires of

Screen Grid (B), thus bypassing the screen grid and continuing to the plate.

ACTUATES RELAYS DIRECTLY!

General Electric's new 7427 cadmium-sulphide photoconductive tube is so sensitive to light variations, and can handle so much current (400 mw max dissipation), that the tube will operate a relay without amplification. Your costs are reduced. Spectrum of the 7427 matches the human eye. Check performance below:



Left: average characteristics, Type 7427

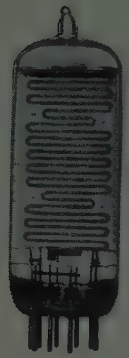
——AC (RMS) operation

- - - - DC operation

Note this new tube's high sensitivity to light, with large current capacity. In series with a relay, the G-E 7427 helps form a simple, economical circuit which will handle scores of lighting, industrial, other control functions.

tubes . New concepts

New materials like 5-ply

like 7427  phototube.

heated cathode in 3DG4.

CUTS HEAT IN TV RECEIVERS!

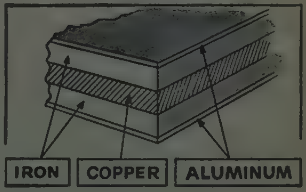
Less heater power...less total power for set...less heat generated! The new General Electric 3DG4 power rectifier tube with direct-heated cathode brings you all three benefits. Special 3-ply cathode requires no filament, teams up with a new high-internal-reflectance plate material for maximum efficiency. Total power required is 42% less than the 5V3. Compare:

	NEW 3DG4	5V3
Heater power	12.5 w	19.0 w
Total watts in tube	29	50
Bulb temperature	171 C	206 C
Output current	350 ma	350 ma

NO "HOT SPOTS" ON ANODES!

General Electric has pioneered the use of 5-ply bonded material for tube anodes. Greatly superior in heat conduction and radiation, the new material prevents the formation of "hot spots" when tubes are running full-load. Gives sustained top-performance capability to a large and growing list of G-E receiving types.

Copper promotes the even distribution and faster dissipation of anode heat. Iron for strength. Aluminum for surface protection.

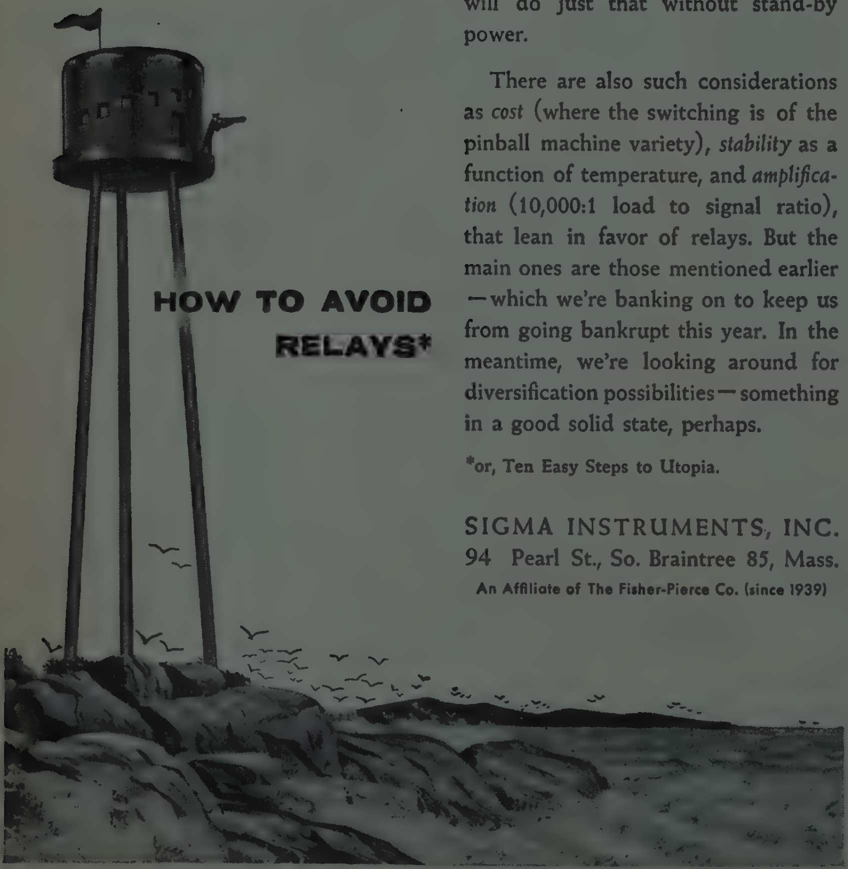


RECEIVING TUBE DEPARTMENT OFFICES:

New York, WI 7-4065, 6, 7, 8....Boston, DE 2-7122....Washington, EX 3-3600....Chicago, SP 7-1600
Dallas, RI 7-4296....Los Angeles, GR 9-7765, BR 2-8566....San Francisco, DI 2-7201.

Progress Is Our Most Important Product

GENERAL ELECTRIC

HOW TO AVOID RELAYS*

If relays cause you as much trouble as they do us, you will undoubtedly welcome information on how to get rid of them. Probably the most fashionable way to do the switching is to use transistors, and as a public service Sigma hereby offers some application data toward this end. The Search for Truth must go on.

Right off the bat, it must be conceded that transistors have the edge in several important physical and dynamic respects. Relays are certainly bigger, heavier and slower, and their useful life is nowhere near infinite—primarily because they all have such old-fashioned things as moving parts. Nor are relays immune to unlimited shock and vibration (the best we've been able to do on a subminiature type, and keep it operating within spec, is 30 g's to 5000 cycles).

There are a few things relays *are* good for, however, even though "Relayized" may never sell a single product. For instance: signal circuits can be isolated from load circuits . . . signal and load can be AC or DC, in any combination . . . circuits with high voltage to ground present no particular problems, and relatively high voltage loads can be handled . . . inductive loads can be switched "off" when they're supposed to be off. On "sliding" or slowly varying signals, the right relay will also provide clean, positive switching and it won't fry if the circuit develops a mild defect. It is true, if not grammatical, to say that a relay is many orders more "off" and several orders more "on" than those other things.

The fact that relay contacts more closely approximate the ideal switch—no ohms one way and infinite ohms the other way—also means something when dry circuit switching is your problem. With loads in the order of 0.1 microwatt, a properly designed relay can provide dependable switching.

Further, if 3-position, polar, center-stable switching (Sigma "Form X") is needed, a single relay will do the job. And if the requirement calls for having the switch "remember" and stay in the last switched position, a polarized, magnetic latching relay (our "Form Z") will do just that without stand-by power.

There are also such considerations as *cost* (where the switching is of the pinball machine variety), *stability* as a function of temperature, and *amplification* (10,000:1 load to signal ratio), that lean in favor of relays. But the main ones are those mentioned earlier—which we're banking on to keep us from going bankrupt this year. In the meantime, we're looking around for diversification possibilities—something in a good solid state, perhaps.

*or, Ten Easy Steps to Utopia.

SIGMA INSTRUMENTS, INC.
94 Pearl St., So. Braintree 85, Mass.
An Affiliate of The Fisher-Pierce Co. (since 1939)

Section Meetings

(Continued from page 84A)

DENVER

"Telebit-Digital Telemetry on Explorer VI," Dr. G. E. Mueller, Space Technology Laboratories. 11/19/59.

"Large Scale Electronic Computers and Scientific Computers," Dr. J. J. Sopka, National Bureau of Standards. 1/21/60.

DETROIT

"Surgical Techniques for Reconstructing the Middle Ear," Dr. H. F. Schuknecht, Henry Ford Hospital; "Psychophysical Studies of Hearing," W. P. Tanner, Jr., University of Michigan Research Institute. 1/8/60.

ELMIRA-CORNING

"Advances in Screen Structure and Data Distribution for the ELF Display System," Dr. E. A. Sack, Westinghouse Electric Corp. 1/11/60.

"Recent Advances in Field Emission," Dr. F. Charbonnier, Linfield Research Institute. 2/1/60.

EL PASO

"Technical Training Is Now Being Automated," Dr. R. F. Mager, Fort Bliss Human Research Section. 1/7/60.

EMPORIUM

"Automation for Quality Control Testing of Electron Tubes," R. A. McNaughton, Sylvania Electronic Tubes. 1/19/60.

EVANSVILLE-OWENSBORO

"Nuclear Radiation & Electronic Equipment," J. R. Crittenden, G.E. Co. 1/20/60.

"The Next Twenty Years in Space," Dr. Leo Steg, G.E. Co. 2/10/60.

FLORIDA WEST COAST

"IRE—A Look at Headquarters," Dr. R. L. McFarlan, IRE President. 2/8/60.

FORT HUACHUCA

"Log Periodic Antennas," Dr. Duhamel, Collins Radio Co. 1/25/60.

FORT WORTH

"Moon Bounce Communications Systems," Dr. A. Straiton, Univ. of Texas. 1/15/60.

GAINESVILLE

"Current Problems in Infrared Physics and Technology," Dr. S. S. Ballard, University of Florida. 2/10/60.

HAMILTON

"The Vidicon and Related Tubes," F. Townsend, Westinghouse Elec. Corp. 1/11/60.

"A Bright Radar Display System," Dr. T. W. East, Raytheon Canada, Ltd. 2/8/60.

HAWAII

"Engineering Aspects of the Integrated Fire and Police Department Communications System," Harold Chun, Honolulu Police Department. 10/14/59.

"Novel Compression and Expansion Methods for Audio and Video Use," W. R. Aiken, Kaiser Engineers. 11/11/59.

"Ionospheric Scatter Systems," J. S. McLeod, Page Communications. 1/13/60.

HOUSTON

Business meeting. 1/19/60.

HUNTSVILLE

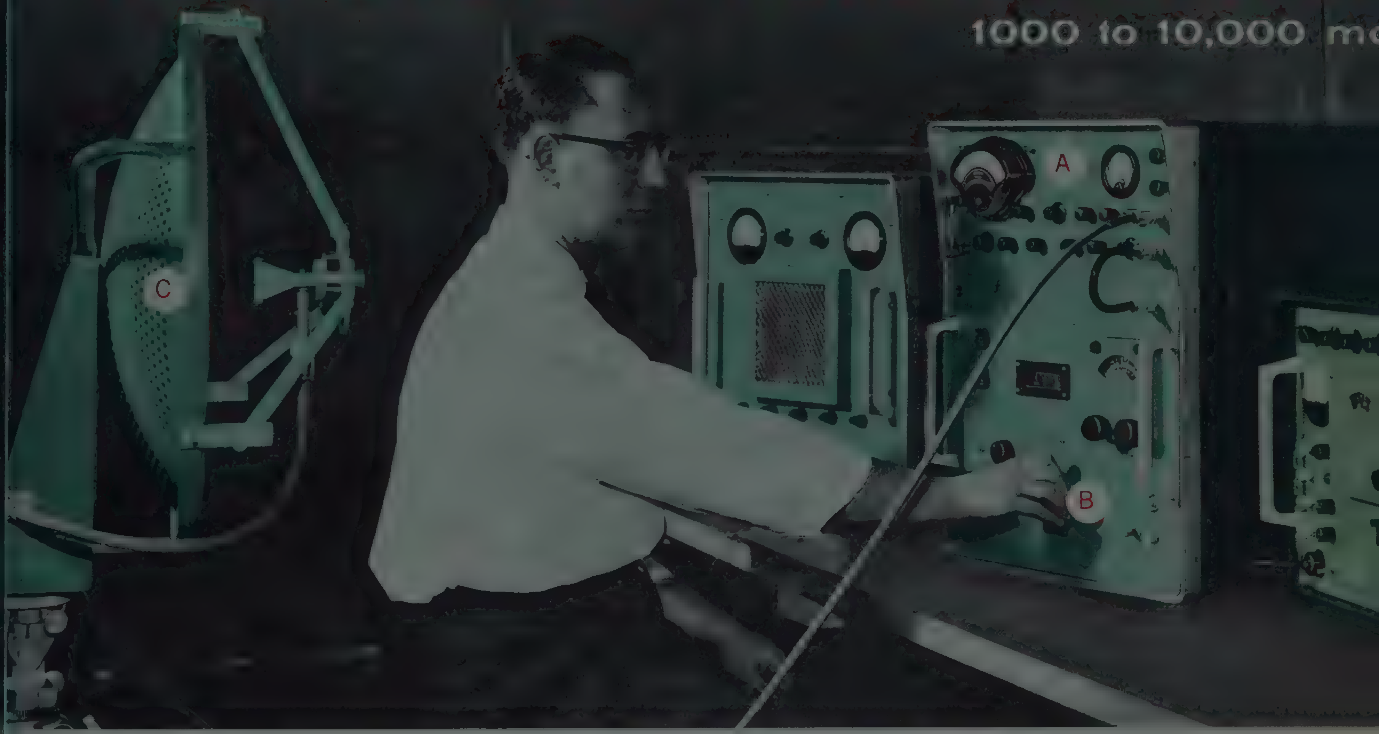
"Pulse Duration Modulation Techniques," Dr. T. G. Antzack, Electro-Mechanical Research, Inc. 1/28/60.

(Continued on page 92A)

Exclusively from Polarad...

CALIBRATED MICROWAVE FIELD INTENSITY RECEIVER SYSTE

1000 to 10,000 mc



Model FIM being used to check radiated interference against MIL-I-26600

The complete integrated Model FIM includes:

- Ⓐ **A monitoring unit.** Meter indicates average, peak or quasi-peak value of r-f signals. Outputs for video, audio and recorder.
- Ⓑ **R-F Plug In Tuning Units.** Four units cover the frequency range 1000 to 10,000 mc. UNI-DIAL tuning control tunes both the receiver and internal signal calibrator.
- Ⓒ **Calibrated Antenna System** complete with tripod and a calibrated omni-directional Broadband antenna or a separate directional antenna to match each R-F tuning unit.

Use this precision instrument for antenna pattern and shielding analysis, r-f power and voltage measurements, determination of field intensity, signal propagation patterns and other microwave parameters.

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☐ Model N-1 Precision Noise Generator (see reverse side of page)

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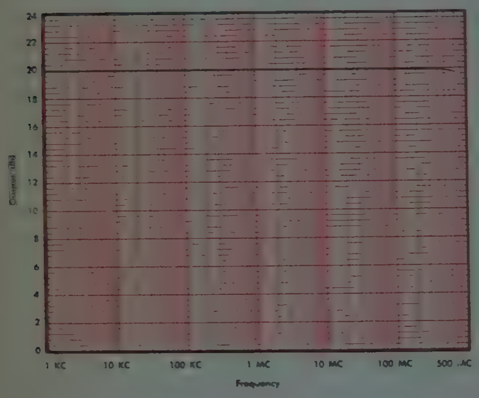
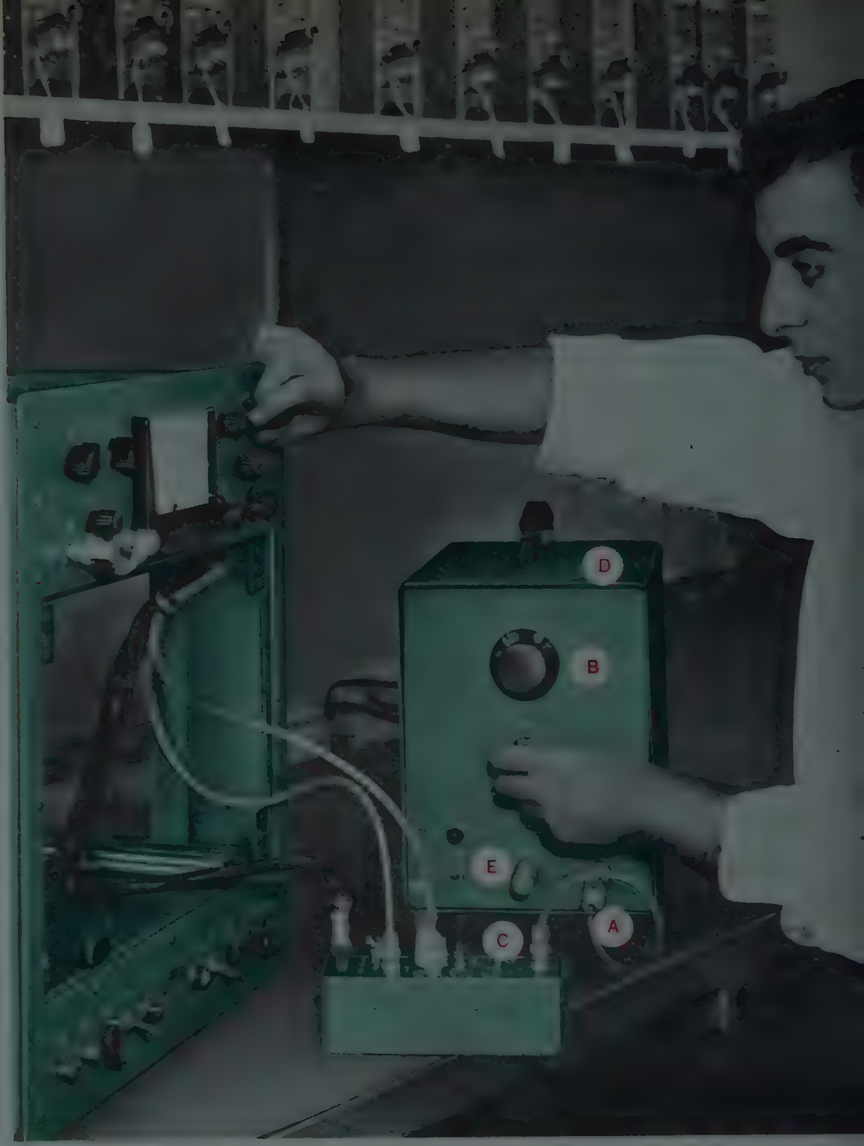
- Ⓐ 1 kc to 500 mc
- Ⓑ 1 to 20 db, adjustable
- Ⓒ 0.25 db output accuracy
- Ⓓ Long-life, low-maintenance noise tubes
- Ⓔ Stable output

Noise output constant over entire frequency range. Maintains set level into any terminating impedance, 50 to 10,000 ohms.





Production line noise figure measurements, made rapidly and accurately with Polarad Model N-1.

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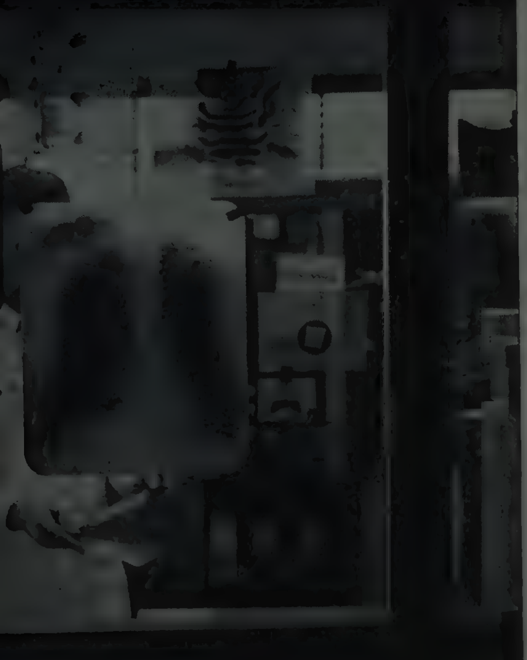


Samarium-145, Samarium-153, Gadolinium-153.
Scientists at the General Motors Research Laboratories began three years ago to measure and re-evaluate the nuclear characteristics of these rare earth isotopes — their half-lives, photon emissions, thermal neutron cross sections.

Conclusion: the radioisotopes had attractive possibilities in industrial and medical radiography, emitting almost pure gamma rays or X-rays (photons) in the low energy range of 30 to 100 kev.

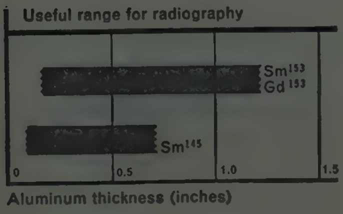
The transition from research to hardware came through two key developments. First, cermet pellets were fabricated using only a few milligrams of the rare earth oxides. Then the irradiated pellets were packaged in special bullet-size holders.

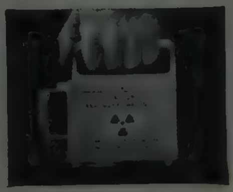
The resulting small, sealed radiographic sources are now being field and laboratory tested. Two excellent applications: "inside-out" checks of hollow shapes inaccessible to X-ray tubes, and radiography of thin steel sections and low density materials such as aluminum or human bone. For example, a recent medical milestone was a chest radiograph of a living person made with a Sm^{153} source. The portable exposure unit to shield the source weighed only 18 pounds.

This isotope radiography program is but one example of the work underway in GM Research's modern isotope laboratory — work that means, through science, "more and better things for more people."

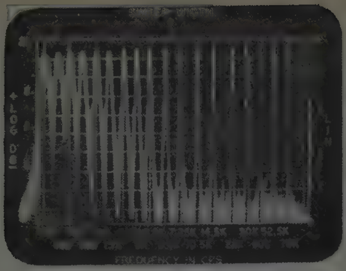
General Motors Research Laboratories

Warren, Michigan





Sm^{153} exposure unit.



TMI-la analysis of 18 IRIG FM/FM channels with slight pre-emphasis.

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Model TMI-1a, 350 cps to 85 kc: for rapid monitoring of all or any part of subcarrier spectrum, checks pre-emphasis, analyzes subcarrier, cross talk, VCO deviations, etc.

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Section Meetings

(*Continued from page 88A*)

ITHACA

"Electronic Beam Parametric Amplifiers," Dr. Robert Adler, Zenith Radio Corp. 1/14/60.

"The Response of Electronic Equipment to Acoustic Noise Fields," N. Doelling, Bolt Beranek & Newman, Inc., 2/11/60

KANSAS CITY

Business meeting. 12/8/59.

"Sound Transmission in Ocean Waters and Some of Its Problems," C. M. Wallis, University of Missouri. 1/12/60.

LITTLE ROCK

"Significance of Noise and Noise Figure in Radio Communication," Alden Packard, AR & T Electronics, Inc. 1/18/60.

"Methods of Characterizing Semiconductor Surfaces," Dr. C. G. Peattie, Texas Instruments, Inc. 2/15/60.

LONDON (ONTARIO)

"The D.R.T.E. Digital Computor," C. D. Florida, Department of National Defence, Ottawa. 1/5/60.

LONG ISLAND

Introduction to Lecture Series. Dr. C. J. Mundo, Jr. Arma Corp.; "Psycho-Acoustics," Dr. W. Rosenblith, MIT. 10/1/59.

"Spectroscopy and Biological Cell Structure," Dr. L. Ornstein, Mt. Sinai Hospital. 10/8/59 (2nd of Lecture Series).

"Neurophysiology of Human Control Processes," Dr. H. Grundefest, Columbia University. 10/15/59 (3rd of Lecture Series).

"Cell Structure and the Electron Microscope," Dr. K. R. Porter, Rockefeller Institute. 10/22/59 (4th of Lecture Series).

"Neurophysiology of Sight," Dr. H. Hartline, Rockefeller Institute. 10/29/59 (5th of Lecture Series).

"Electronic and Computer Techniques in Diagnosis of Heart Diseases," W. E. Tolles, W. Carbery, C. Steinbery, Airborne Instruments Laboratory. (6th of Lecture Series). 11/5/59.

"Strategy of Deterrence," Harry Davis, Rome Air Development Center. 1/12/60.

LOS ANGELES

"Aerospace Power for National Security," Gen. O. J. Ritland, ARDC. 2/4/60.

LOUISVILLE

Tour of Standiford Field Radar Flight Control Center with comments by the F.A.A. Staff. 8/27/59.

"WAVE's New Building and Facilities," Wilbur Hudson, WAVE. 9/24/59.

"Jet Aircraft Control and Flight Simulator," V. M. Yahne, Kentucky Air National Guard. 10/22/59.

"Lighting of Radio and Television Towers," O. W. Towner, WHAS. 11/20/59.

LUBBOCK

"Interference at a Common Tower," David Land, Motorola. 1/26/60.

MIAMI

"Electronic Education and Engineering in India," Dr. P. H. Craig, U. S. State Department. 1/15/60.

MILWAUKEE

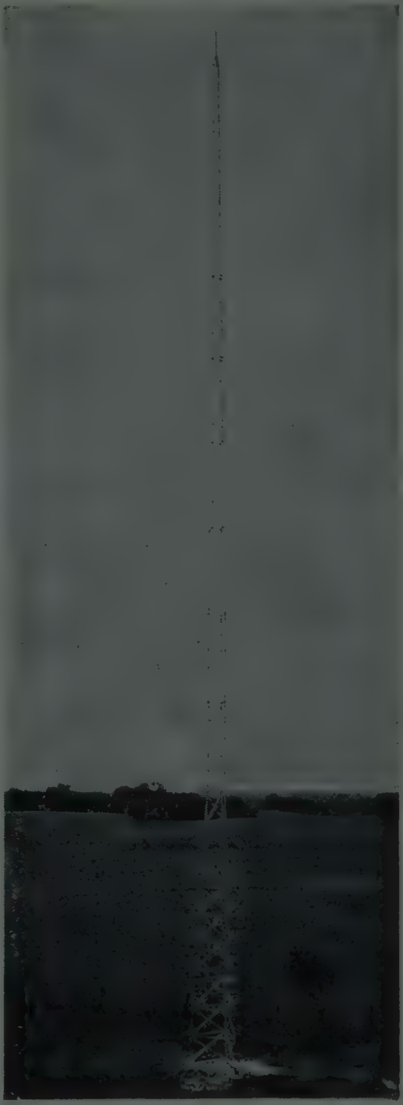
"Radiometry," Ted Seling, A. C. Spark Plug Co. Presentation of Fellow Award. 1/20/60.

MONTREAL

"Measured Retaliation," M. Jennings, Raytheon Corp. 1/20/60.

(*Continued on page 94A*)

New ROHN
SELF SUPPORTING
COMMUNICATION TOWER



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HIGHER CURRENT THAN EVER BEFORE FOR MILITARY AND COMMERCIAL USE

	2N1518	2N1519	2N1520	2N1521	2N1522	2N1523
Maximum Collector Current (Amps.)	25	25	35	35	50	50
Maximum Collector to Base Volts, Emitter Open, Max Ico 4ma	50	80	50	80	50	80
Minimum Open Base Volts (1-Amp. Sweep Method)	40	60	40	60	40	60
Maximum Saturation Volts at Maximum Collector Current	0.7	0.7	0.6	0.6	0.5	0.5
Gain at Ic at 15 Amps.	15-40	15-40	17-35	17-35	22-45	22-45
Minimum Gain at Maximum Collector Current	12	12	12	12	12	12
Thermal Resistance Junction to Mounting Base (°C/Watt)	0.8	0.8	0.8	0.8	0.8	0.8

Characteristics at 25°C   Maximum Junction Temperature 95°C

A new family of high current transistors featuring the 50-ampere 2N1522 and 2N1523. Two 25- and two 35-ampere types round out the line. All thoroughly tested and completely reliable. Available in production quantities. Call or write your nearest Delco Radio sales office for full product information and applications assistance.

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Noise spectrum analysis using internal video smoothing filter. Noise envelope average vs. frequency seen in readily interpreted plot. Internal marker pips are 500kc apart.

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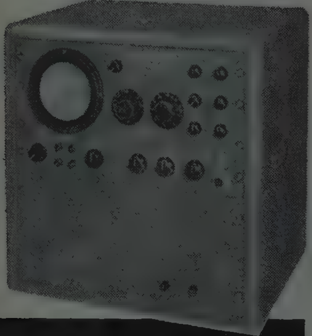

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• 0-3mc sweep width continuously calibrated • Variable center frequency control cailibrated from 0 to 23.5 mc • Adjustable resoultion; selectivity, 200cps to 30kc • Variable sweep rate 1cps to 60 cps • Lin, 40db log and square law amplitude scales • High sensitivity: 20uv full scale • Calibrated 100 db input attenuators

Also available: Model SPA-3, 200cps-15mc. Same at SPA-3/25, except variable center frequency control is calibrated 0 to 13.5mc.

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Section Meetings

(Continued from page 92A)

NEWFOUNDLAND

"Ladies Nite," Dinner Meeting and Dance. 12/14/59.

"Supervisory Control of Hydro-Electric Plants," M. Gladden, Newfoundland Light & Power Co. 1/21/60.

NEW YORK

"Wall Street Analyst's Look at Electronics," Discussion by Panel consisting of G. Edgar, M. Loeb; M. Heilbrun, Rhoades & Co.; A. Cohn, Bache & Co.; Miss P. Shulder, Merrill Lynch, Pierce, Fenner & Smith, Inc. 12/2/59.

NORTH CAROLINA

"Sampling Techniques for Automatic Waveform Analysis," M. R. Sundquist, Bell Telephone Labs. Presentation of Fellow Award. 1/15/60.

"Subscriber Line Data Transmission Systems," P. J. Grunfelder, Western Electric Co., Inc. 2/19/60.

NORTHERN ALBERTA

"Filters and Duplexers," Dr. W. V. Tilston, Sinclair Radio Labs. 1/19/60.

NORTHWEST FLORIDA

"Recent Biological Experiments in Ballistics Missiles," Dr. D. E. Stullken, NATC, Pensacola. 12/15/59.

OKLAHOMA CITY

Tour of Amplitron Installation and Air Navigation Facilities, Will Rogers Field, Oklahoma City. 1/12/60.

"Use of Transistors in Seismic Work," J. D. Delbridge, Shell Oil Co. 2/9/60.

ORLANDO

"Current Problems in Infrared Physics & Technology," Dr. S. S. Ballard, University of Florida; Film: "Bomarc Weapon System." 1/13/60.

"Considerations in Design and Use of Direct Writing Recording Systems," Dr. A. Miller, Sanborn Co. 1/20/60.

"Facts About IRE Headquarters—Present Activities and Future Plans," Dr. R. L. McFarlan, IRE President. 2/1/60.

Seminar on "The Role of a Nuclear Reactor in a University," Dr. T. J. Thompson, MIT; Dr. W. Kerr, University of Michigan; W. J. Clapp, Florida Power Corp.; Dr. W. B. Harrison, Georgia Tech.; Dr. Glenn A. Greathouse, University of Florida; Dr. J. A. Wethington, University of Florida. Presiding was Dr. Walter H. Zinn, General Nuclear Engineering Corp. 2/13/60.

OTTAWA

"Design Problems in Developing Oscilloscopes," M. B. Crouch, Tektronix, Inc. 1/14/60.

Five different subjects by University Students. Presentation of Fellow Award. 2/4/60.

PHILADELPHIA

"Design of the UNIVAC-LARC System," J. P. Eckert, H. Lukoff and A. B. Tonik of Remington Rand. 1/6/60.

PHOENIX

"Signal-Flow Graphs in Sampled-Data Systems," Dr. J. M. Salzer, Ramo-Wooldridge Corp. 1/19/60.

PITTSBURGH

"Instrumentation—Past and Present," R. D. Wyckoff, Gulf Research and Development Corp. 10/5/59.

"The Systems Approach to Molecular Electronics," Dr. A. W. McCourt, Westinghouse Air Arm Division. 10/19/59.

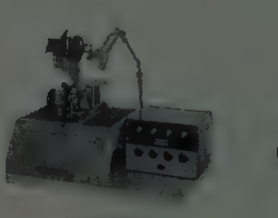
Field tour of Greater Pittsburgh Airport Air Traffic Control Center. 11/9/59.

"Recent Developments in Thermoelectric Cool-

(Continued on page 96A)

BOESCH

subminiature toroidal coil winders



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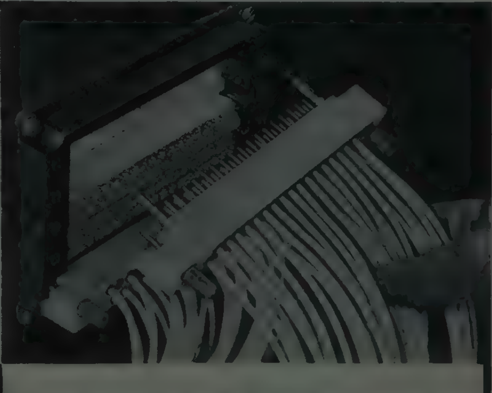
feed-thru, multiple insert

Makes possible the design of lighter and more compact equipment. Each insert holds 35 contacts. Frames available for 5 or 8 inserts.



BURNDY

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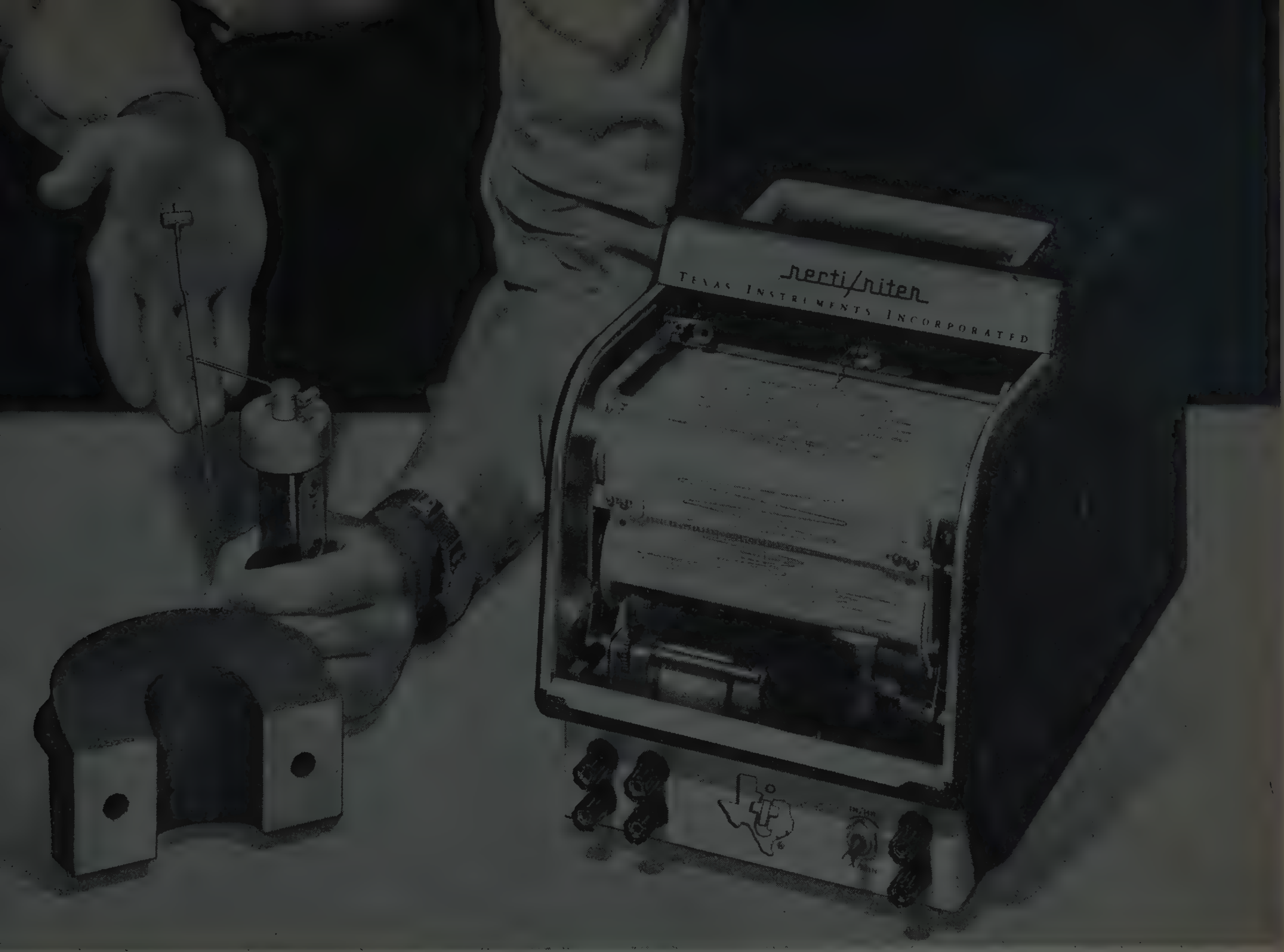

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What simpler and more reliable actuating device can you employ in an amperage-voltage-frequency recording instrument than a d'Arsonval galvanometer . . . a trouble-free horseshoe magnet and a coil of wire? The same is true of the exclusive "recti/rite"® system . . . a simple, shock resistant trigonometric linkage that straightens the arc described by the galvanometer metering arm, changing curvilinear motion to rectilinear motion.

All the other "recti/riter" recorder features which contribute to this instrument's multi-industry acceptance and hardworking reliability are equally simple: The optional a-c or d-c drives couple directly with chart speed change gears to allow ten chart speeds; all routine operations and adjustments are performed "up front"; the non-corrosive, honed metal alloy pens, closed ink system, and large capacity ink well give you long, consistent writing performance.

With all their simplicity and reliability, "recti/riter" recorders are offered in extremely wide and useful Basic Recorder Ranges (Dual channel recorders offer combination of any two ranges):

Two Cycle Pen Response

D-c Milliampere Ranges	¼ ma to 100 ma
A-c Ampere Ranges	0.25 A to 25 A
D-c Ampere Range	100 mv for use with standard shunts
Expanded Scale A-c Voltage Ranges	80-130 V, 160-260 V, 320-520 V
A-c and D-c Voltage Ranges	10 V to 1000 V
Frequency Ranges	50, 60, 400 cps

Five Cycle Pen Response

D-c Milliampere Range	5 ma

Ask the TI engineer about *customized* recorders for your OEM applications. Don't settle for any recorder until you know all the facts on the complete "recti/riter" recorder line.

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new rapid tests of SSB transmissions with ONE compact multi-purpose spectrum analyzer

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SSB-3

. . . simple
. . . versatile
. . . low-priced

Now, in one convenient package, all the equipment you need to set up, adjust, monitor, trouble-shoot SSB and AM transmissions!

- 60 db dynamic range
- 60 cps hum sidebands measurable to —60 db
- Stable tuning head with 2 mc to 40 mc range with direct reading dial
- Sensitive spectrum analyzer with pre-set sweep widths of 150, 500, 2000, 10,000 and 30,000 cps with automatic optimum resolution
- Continuously variable sweep width up to 100 kc
- Two-tone generator with separate audio oscillators with independent frequency and amplitude controls. Output 2 volts max. per tone into 600 ohm load
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Section Meetings

(Continued from page 94A)

ing Devices and Power Generation," Dr. S. J. Angello, Westinghouse Electric Corp. 12/7/59.

"The New Sound"—A special non-technical non-commercial demonstration of new and old techniques for recording by a special RCA demonstration team. 12/14/59.

PORTLAND

Tour of Tektronix Plant. 12/17/59.

Demonstration Lecture on the ESIAC Algebraic Computer, by Dr. M. Morgan, Electro-Measurements, Inc. 1/23/60.

"Time Division in Communication," K. Searcy, Pacific Telephone & Telegraph Co. 2/10/60.

Panel discussion on "An Impedance Plethysmograph-Instrumentation and Clinical Application" and "A Myograph, Instrumentation and Clinical Application," Panel members: Loren Park, Dr. R. J. Underwood, Dr. R. E. Rinehard, and Dr. D. Michael. 2/11/60.

QUEBEC

"The Case for Special Quality Tubes," D. S. Simkins, Rogers Electronic Tubes and Components. 1/27/60.

REGINA

Conducted tour of Regina Airport Radar Installation. 11/27/59.

ROME-UTICA

"Interface Problems in System Design," A. Kunze, Rome Air Development Center. 1/26/60.

SALT LAKE CITY

"A New Look At Relativity," John Anderson, Retired. 11/3/59.

"Traveling Wave Tube Oscillograph," Dr. R. Grow, University of Utah. 12/8/59.

"Pulse Doppler Radar," A. W. Vodak, Sperry Utah Engineering Lab. 1/14/60.

SAN ANTONIO-AUSTIN

"Parametric Amplifiers," Dr. K. Kotzebue, Texas Instruments. 10/16/59.

"Electronic Circuitry for Ultrasonic Application," W. A. Gunkel, Southwest Research Institute. 11/20/59.

"Maximizing Information Rate in Data Systems," Dr. W. A. Youngblood, University of Texas. 12/12/59.

"Techniques of Audio Frequency Measurements," Hoyt Foster, Kelly AFB. 12/18/59.

Presentation of Fellow Award. 1/22/60

SAN DIEGO

"Electroluminescence," S. H. Boyd, Stromberg-Carlson. 1/6/60.

"Optimization of Airborne Special-purpose Computers," R. Williamson, Librascope. 2/3/60.

SCHENECTADY

"Digital Shaft Encoders," R. Knapp, RPI. 1/21/60.

"The Scavenger Amplifier," Dr. W. K. Volkers, Millivac Instrument Corp. 2/9/60.

SHREVEPORT

"Microwave Communications," A. K. Urschel and Mr. Davis, Collins Radio. 1/5/60.

"Transistor Characteristics and Circuits," Prof. E. W. E. Owen, Louisiana State University. 2/2/60.

SOUTH BEND-MISHAWAKA

"Doppler Navigation Radar," R. E. Tollefson, Collins Radio. 9/24/59.

"The IRE in the Space Age," Dr. Ernest Weber, IRE President. 10/14/59.

Tour of RCA Whirlpool Research Laboratories with comments by Dr. D. Cutler. 10/29/59.

"Magnetics—Key to Modern Control," E. V. Wier, Magnetics, Inc. 12/3/59.

(Continued on page 98A)

Wide Band Amplifier Model 530

GENERAL DESCRIPTION

The Model 530 Wide Band Amplifier has been designed to fill a need for an amplifier used principally for voltage amplification of CW or pulsed signals.

This model has many applications in the laboratory and also for television distribution systems. It may be used to increase the output from signal sources within its frequency range, and its bandwidth of 300 mc's makes it ideal for amplifying millimicrosecond pulses.

PRICE $330.

SPECIFICATIONS

Bandpass	10 KC to 300 MC
Voltage Gain	18 db
Input Impedance	135 ohms
Output Impedance	150 ohms
Max. Output Power (into Matched Load)	.08W
Max Output Voltage (into Matched Load)	3.5 Vrms
Max. Peak Pluse Output	7 V pos. or neg.
Rise Time	Less than 2x10^{9} sec
Dimensions	19" front panel—16" wide, 9" deep 3½ high. Power supply included.
Gain Control	Provided on front panel
Tube Complement	Two cascaded stages of eight 6AK5

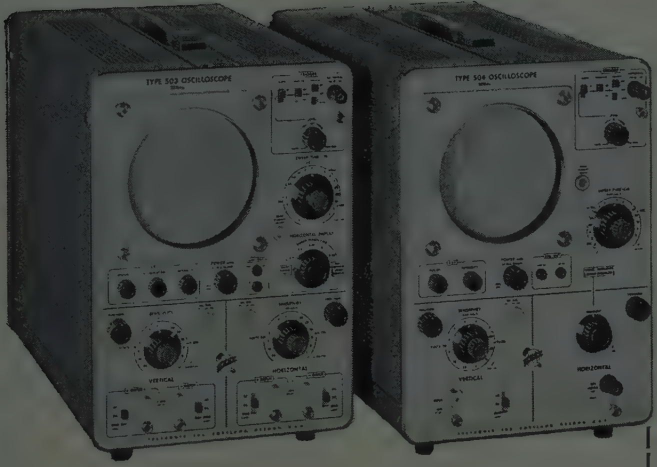
INSTRUMENTS FOR INDUSTRY, Inc.
101 New South Road, Hicksville, L. I., N.Y.

INSTRUMENTS ifi FOR INDUSTRY

Graduate engineers with two or more years of circuit application in the fields of electronics or physics are invited to meet with Mr. John Hicks in an informal interview or send complete resume to: Dir. Personnel, IFI, 101 New South Road, Hicksville, New York.

2 NEW

GENERAL-PURPOSE OSCILLOSCOPES introduce TEKTRONIX QUALITY to the DC-to-450 KC RANGE



The Tektronix Type 503 and Type 504 are the first of a family of new oscilloscopes for the DC-to-450 KC application area.

- **Both feature high reliability, simple operation, light weight.**
- **Each excels in performance characteristics in its class.**
- **Both now established as production instruments.**

Prices

TYPE 503 $625

TYPE 504 525

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Tektronix, Inc.

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Phone MItchell 4-0161 • TWX—BEAV 311 • Cable: TEKTRONIX

TYPE 503

The Type 503 is a differential-input X-Y oscilloscope with the additional features—linear sweeps, dependable triggering, sweep magnifier, bright trace, amplitude calibrator—desirable for general-purpose applications.

FREQUENCY RESPONSE
dc to 450 kc

VERTICAL AND HORIZONTAL AMPLIFIERS
Differential input at all attenuator settings.
1 mv/cm to 20 v/cm in 14 calibrated steps.
Continuously variable between steps, and to approximately 50 v/cm uncalibrated.
Constant input impedance at all sensitivities (standard 10X probes can be used).

SWEEP RANGE
1 μsec/cm to 5 sec/cm in 21 calibrated steps.
Sweep time adjustable between steps, and to approximately 12 sec/cm uncalibrated.

SWEEP MAGNIFICATION
X2, X5, X10, X20, and X50 Magnification.

AMPLITUDE CALIBRATOR
500 mv and 5 mv peak-to-peak square-wave voltages are available from front panel.

3-KV ACCELERATING POTENTIAL
5-inch Tektronix crt provides bright trace, 8-cm by 10-cm viewing area.

EASY TRIGGERING
Fully automatic, amplitude-level selection on rising or falling slope of signal, or free-run (recurrent). AC or DC coupling, internal, external, or line.

REGULATED POWER SUPPLIES
All critical dc voltages electronically-regulated, plus regulated heater supplies for the input stages of both amplifiers.

SIZE AND WEIGHT
13½" h, 9¾" w, 21½" d—approximately 29 lbs.

TYPE 504

The Type 504 has the basic features desirable for most general-purpose applications — sensitive vertical amplifier, linear sweeps, easy triggering, amplitude calibrator.

FREQUENCY RESPONSE
dc to 450 kc

VERTICAL AMPLIFIER
5 mv/cm to 20 v/cm in 12 calibrated steps.
Continuously variable between steps, and to approximately 50 v/cm uncalibrated.
Constant input impedance at all sensitivities (standard 10X probe can be used).

SWEEP RANGE
1 μsec/cm to 0.5 sec/cm in 18 calibrated steps.
Sweep time adjustable between steps, and to approximately 1.2 sec/cm uncalibrated.

AMPLITUDE CALIBRATOR
500 mv and 25 mv peak-to-peak square-wave voltages are available from front panel.

HORIZONTAL INPUT
0.5 v/cm, with variable attenuator.

3-KV ACCELERATING POTENTIAL
5-inch Tektronix crt provides bright trace, 8-cm by 10-cm viewing area.

EASY TRIGGERING
Fully automatic, amplitude-level selection on rising or falling slope of signal, or free-run (recurrent). AC or DC coupling, internal, external, or line.

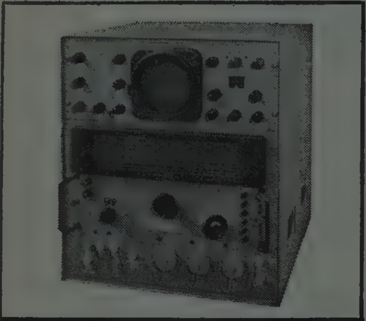
REGULATED POWER SUPPLIES
All critical dc voltages electronically-regulated, plus regulated heater supplies for the input stages of the vertical amplifier.

SIZE AND WEIGHT
13½" h, 9¾" w, 21½" d—approximately 29 lbs.

Rack-mounting models will be available, of course!

TEKTRONIX FIELD OFFICES: Albuquerque, N. Mex. • Atlanta, Ga. • Baltimore (Towson, Md.) • Boston(Lexington, Mass.) • Buffalo, N.Y. • Chicago (Park Ridge, Ill.) • Cleveland, Ohio • Dallas, Texas • Dayton, Ohio • Denver, Colo. • Detroit (Lathrup Village, Mich.) • Endicott (Endwell, N.Y.) • Greensboro, N.C. •Houston, Texas • Kansas City (Mission, Kan.) • East Los Angeles, Calif. • West Los Angeles, Calif. • Minneapolis, Minn. • New York City Area (Albertson, L.I., N.Y. • Stamford, Conn. • Union, N.J.) • Orlando, Fla. • Philadelphia, Pa. • Phoenix, (Scottsdale, Ariz.) • San Diego, Calif. • San Francisco (Palo Alto, Calif.) • St. Petersburg, Fla. • Syracuse, N.Y. • Toronto (Willowdale, Ont.) Canada • Washington, D.C. (Annandale, Va.)

TEKTRONIX ENGINEERING REPRESENTATIVES: Hawthorne Electronics, Portland, Oregon • Seattle, Washington. Tektronix is represented in twenty overseas countries by qualified engineering organizations.



extreme sensitivity 10 mc to 44,000 mc

PANORAMIC'S SPA-4 SPECTRUM ANALYZER

MORE USEABLE SENSITIVITY

BAND	RF SENSITIVITY*
10 — 420 MC	—95 to —105 dbm
350 — 1000 MC	—90 to —100 dbm
910 — 2200 MC	—90 to —100 dbm
1980 — 4500 MC	—80 to — 90 dbm
4.5 — 10.88 KMC	—80 to — 95 dbm
10.88 — 18.0 KMC	—70 to — 90 dbm
18.0 — 26.4 KMC	—60 to — 85 dbm
26.4 — 44.0 KMC	—55 to — 85 dbm

*measured when signal and noise equal 2X noise

Using one tuning head which contains one triode and two Klystron oscillators, Model SPA-4 offers more exclusive advantages for applications demanding extreme sensitivity, stability, versatility, accuracy.

- Three precisely calibrated amplitude scales—40 db log, 20 db linear, 10 db power.
- Two independent frequency dispersion ranges—continuously adjustable—0-70 mc and 0-5 mc. Negligible internal frequency modulation permits narrow band analysis of FM problems.
- Variable I.F. bandwidth from 1 kc to 80 kc.
- Push-button frequency selector.
- Synchroscope output with 40 db gain.
- Accurate measurement of small frequency differences. A self-contained marker oscillator, modulated by a calibrated external generator, provides accurate differential marker pips as close as 10 kc.

Tremendous flexibility and many unique advances of Panoramic's compact SPA-4 make it unsurpassed for visually analyzing FM, AM and pulsed signal systems; instabilities of oscillators; noise spectra; detection of parasitics; studies of harmonic outputs; radar systems and other signal sources.

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Cables: Panoramic, Mt. Vernon N. Y. State

Section Meetings

(Continued from page 96A)

SOUTH CAROLINA

"The Development of the Modern Electron Tube," Dr. W. R. Ferris, University of South Carolina, 2/5/60.

SYRACUSE

"Centralized Railroad Traffic Control," C. F. Mulranen, N.Y. Central R.R. 1/21/60.

TOLEDO

"Tomorrow's Communications," D. F. Jones and R. F. Palomo, Ohio Bell Telephone Co. 1/14/60.

TUCSON

"Log Periodic Antennas," Dr. R. H. Du Hammel, Collins Radio Co. 1/26/60.

TULSA

"A Low Phase Distortion Seismic Amplifier," Dr. H. B. Ferguson, Jersey Production Research Co. 1/21/60.

VIRGINIA

"High Power Radiation Hazards (HERO)," J. N. Payne, U. S. Naval Weapons Lab. 1/8/60.

WASHINGTON

"Integrated Electronics," Dr. D. A. Jenny, RCA Labs. 1/26/60.

"How Are We Conserving Our Natural Resource—The Radio Spectrum?" panel discussion with Commodore E. M. Webster, Moderator. Panel members: T.A.M. Craven, FCC; Capt. P. Miles, Office of Civil Defense & Mobilization; W. H. Watkins, FCC. 2/1/60.

WILLIAMSPORT

"English Contributions to Television," F. Rutter, Westinghouse Electronics. 1/27/60.

"Propulsion for Space Flight," Dr. E. Petrick and H. Szymanowski, Curtiss-Wright Corp. 2/1/60.

WINNIPEG

"Transient Response from Frequency Response by Digital Computer Methods," M. Ito, University of Manitoba; "Mercury Arc Rectifiers, Rectifier Transformers and Rectification Systems," M. Mayer, University of Manitoba. 12/2/59.

"Antenna Research," Dr. W. V. Tilson, Sinclair Radio Lab. 1/20/60.

SUBSECTIONS

BUENAVENTURA

"Some Aspects of Instrumentation for Ballistic Missiles and Other Rockets," John Masterson, Pacific Missile Range; "Closed Circuit TV and Swept Band Transmission Measuring Set Demonstrations," S. Wilson, Hallamore Electronics Co. 1/13/60.

EAST BAY

"High Speed Data Reduction," J. Russell, Lawrence Radiation Lab. 10/5/59.

EASTERN NORTH CAROLINA

"Soil Moisture Content by Neutron Scattering," W. F. Troxler, Troxler Electrical Lab. Election of officers. 1/8/60.

"Semi-conductor Charged-particle Detectors," J. L. Blankenship, Oak Ridge National Lab. 2/12/60.

FAIRFIELD COUNTY

"Technical Manpower, A Vital Resource," Dean A. H. Jacobson, Pennsylvania State University. 1/21/60.

KITCHENER-WATERLOO

"Reproduction of Heart Sounds," L. M. Steinberg, University of Toronto. 1/25/60.

(Continued on page 102A)

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Kodak reports on:

and and . . . sunshine for the plastics man who worries . . . organic chemistry and optical pumping . . . the question of whether or not instrumentation people really need ultra-fast film

Infrared detectors for sale

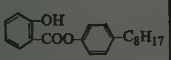
Our involvement in infrared detectors has deepened.

Pictured at the top of this page are, respectively, the simplest kind of *Kodak Ektron Detector* with a rectangular sensitive area of any reasonable dimensions, available either in a 3-pin miniature cable socket or unmounted with resolderable leads; an "immersed" detector with detecting substance deposited on the plano surface of a radiation-collecting lens; a detector mounted in a Dewar for cooling by cryostat. We also deposit detecting substance in separated or intricate configurations as ingenuity, under necessity's goad, may provide.

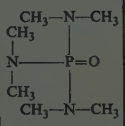
Since any of these physical forms can now be provided with any of six different kinds of lead sulfide or lead selenide depositions, as governed by spectral sensitivity, response time, temperature, and ambient humidity, the print gets quite fine in a folder we are publishing this month to guide the selection of *Kodak Ektron Detectors.*

A free copy is available from Eastman Kodak Company, Apparatus and Optical Division, Rochester 4, N. Y. It is designed to make the sale with minimum further correspondence. To give you an idea, the off-the-shelf, one-only price scale starts from $14.50.

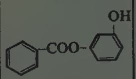
Corruption can be arrested



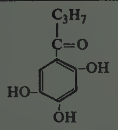
OPS will arrest corruption



HPT
will arrest corruption



RMB
will arrest corruption



THBP will arrest corruption

When plastics stand up to cracking, crazing, embrittlement, gumming, and discoloration with the passage of time and sunshine, it's OK to credit the integrity of the maker's name.

The concept of integrity is more readily grasped than is the principle that every C-C bond, every O-H bond, every C-Cl bond, every C=N bond, has its own price—a photon of ultraviolet radiation just right in energy to snap it. Thus corruption of plastics, a chain reaction like other corruptions, begins. A rival substance that grabs off the u-v photons first and degrades their energy will delay the corruption for a long time.

Other corruptions, initiated by atmospheric oxygen, can be delayed by compounds that stop the action by supplying hydrogen at a critical juncture. Of course, choice in inhibitors is greatly restricted by many considerations of physical and chemical compatibility.

Obviously, we think we have made a good choice in these new ones:

Eastman Inhibitor OPS (p-Octylphenyl Salicylate), at 1 to 5% levels, a preventive of the photo-oxidation that forms deleterious carbonyl groups in polyethylene and polypropylene.

Eastman Inhibitor HPT (Hexamethylphosphoric Triamide), very pale, water-soluble, a liquid u-v inhibitor for use with a heat stabilizer in poly(vinyl chloride) at 1 to 2%.

Eastman Inhibitor RMB (Resorcinol Monobenzoate), a noncoloring u-v inhibitor at 1% levels for transparent cellulose acetate butyrate. Superior to phenyl salicylate. Also protects cellulose acetate, cellulose acetate propionate, polystyrene, poly(vinyl chloride), and ethylcellulose.

Eastman Inhibitor THBP (2,4,5-Trihydroxybutyrophenone), an antioxidant for polyolefins and various paraffin waxes. May be added in solid form directly during the milling process, or in solvent solutions to dry powders or melts, at 0.01 to 0.1% concentrations. Especially valuable for absence of stain at the low concentrations.

Isn't it fun how in the chemical industry practically everybody is simultaneously practically everybody else's customer, supplier, and competitor! Data sheets and development samples (for those in a position to evaluate them) from Eastman Chemical Products, Inc., Kingsport, Tenn. (Subsidiary of Eastman Kodak Company).

Man passing out dotriacontane

The atomic physicists are looking for substances against which polarized free electrons can bounce without having their spins inverted. The physicists want to coat a substance like that on the walls of evacuated glass vessels. It should have no free or unpaired electrons, no crystal structure. To an electron scooting by, such a molecule should appear magnetically inert. The small residual pressure tolerable in the vessel should be mostly of rubidium vapor and not of the non-inverting substance. Also the coating must have good transmittance at around the 8000A wavelength of the radiation fed to excite the rubidium.

"Optical pumping," the theme these experiments share, has to do with the manipulation of quantum levels so that emission from the excited alkali metal atoms can be tuned with great elegance. Pressure to improve understanding of the phenomena comes from the need for instrumentation to study the Van Allen radiation belts and for satellite-borne atomic clocks to check out Einstein's old predictions about the meaning of time as told from a moving timepiece.

Some little samples of non-inverting substance were passed out at a recent optical pumping conference by a man who may well have bought the material from us as *Dotriacontane* (Eastman 3555). Spermacetti wax once had a role in physics as the material of the standard candle which defined candlepower. When chemists attacked it, they found cetyl palmitate, $C_{16}H_{33}OCOC_{15}H_{31}$. We have to make a long, uniform, inert, homogeneous molecule out of that. In effect, the two oxygens the whale put in the middle of the molecule are replaced by hydrogens. That is *Dotriacontane*. The climax, if any, will come when word filters through that it has done the job.

The moral is that with some 3800 organic compounds to milk, we need never be at a loss for words. They're all stocked by Distillation Products Industries, Eastman Organic Chemicals Department, Rochester 3, N. Y. (Division of Eastman Kodak Company).

1600—no waiting

How come after all those promises we have made to innumerable instrumentation people over the years that some day there would be 16mm, 35mm, and 70mm film as fast as *Kodak Royal-X Pan Recording Film* now is—Index 1600—how come we now find ourselves in the ridiculous position of being able to make it at a greater rate than they're buying it? How come?

Don't they know that a note or phone call to Eastman Kodak Company, Photo Recording Methods Division, Rochester 4, N. Y., will set up the channel to supply it through a local dealer?

Price quoted is subject to change without notice.

This is another advertisement where Eastman Kodak Company probes at random for mutual interests and occasionally a little revenue from those whose work has something to do with science

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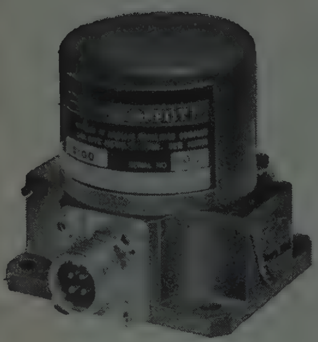

ELECTROHYDRAULIC SERVO VALVE

Kearfott's unique approach to electrohydraulic feedback amplification design has resulted in a high-performance miniature servo valve with just two moving parts. Ideally suited to missile, aircraft and industrial applications, these anti-clogging, 2-stage, 4-way selector valves provide high frequency response and proved reliability even with highly contaminated fluids and under conditions of extreme temperature.

TYPICAL CHARACTERISTICS

Quiescent Flow	0.15 gpm
Hysteresis	3% of rated current
Frequency Response	3 db @ 100 cps
Supply pressure	500 to 3000 psi
Temperature-Fluid & Ambient	−65° F to +275° F
Flow Rate Range	.3 to 10 gpm
Weight	10.5 ounces

Write for complete data.

BASIC
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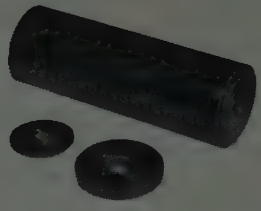
FROM KEARFOTT



FERRITES

Kearfott's Solid State Physics Laboratory formulates, fires and machines permanent magnet ferrite materials of various compositions. Typical high-efficiency array utilizes Kearfott PM-3 ferrite material with specially designed pole pieces to produce a design both smaller and lighter than other arrays of equivalent magnetic field strength. Because magnets may be custom engineered to specific requirements, user is not restricted to stock magnet types, thereby providing greater latitude in parameters for focusing arrays. Pole pieces may also be provided according to specification, with the added assurance that, because of special Kearfott design techniques, B axial magnetic fields approximately 10% higher than those generally obtained in standard types may be produced.

TYPICAL CHARACTERISTICS

Peak Magnetic Field Strength	1200 gauss
Period	0.560 in.
Length	5.64 in.
Inside Diameter of Pole Pieces	0.400 in.
Outside Diameter	2.0 in.
Weight	3.2 pounds

Write for complete data.

BASIC
BUILDING
BLOCKS
FROM KEARFOTT

INTEGRATING TACHOMETERS

Kearfott integrating tachometers, special types of rate generators, are almost invariably provided integrally coupled to a motor. They feature tachometer generators of high output-to-null ratio and are temperature stabilized or compensated for highest accuracy integration and rate computation. Linearity of these compact, lightweight tachometers ranges as low as .01% and is usually better than ± .1%.

TYPICAL CHARACTERISTICS

	Size 11 (R660)
Excitation Voltage (400 cps)	115
Volts at 0 rpm (RMS)	.020
Volts at 1000 rpm (RMS)	2.75
Phase shift at 3600 rpm	0°
Linearity at 0-3600 rpm	.07
Operating Temperature Range	−54° +125°

Write for complete data.

Miniature Floated Gyro

Precise Angle Indicator

20 Second Synchro

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Engineers: Kearfott offers challenging opportunities in advanced component and system development.

TYPE	SIZE	RESISTANCE (ohms)	WATTAGE	TC
NF **Fusion sealed.** Glass encapsulation is fusion sealed to leads and has zero moisture absorption. Exceeds requirements of MIL-R-10509C, Char. B.	NF60 NF65	100 to 100 K 100 to 360 K	⅛ @ 70° C. ¼ @ 70° C.	±.03%/° C. from −55° C. to +150° C., ref. to 25° C.
N **Fixed film.** Extremely low noise level. 0.1 microvolt/volt. Derating to 140° C. Average resistance change after 5000 hrs. is less than 1%. Exceeds MIL-R-10509B, Char. X specs.	N20 N25 N30	10 to 500 K 10 to 1.5 meg. 30 to 4.2 meg.	½ @ 40° C. 1 @ 40° C. 2 @ 40° C.	±.03%/° C. from −55° C. to +105° C., ref. to 25° C.
S **Fixed film: high temperature.** Less than 0.35% resistance change after 1000 hrs. of load-life tests at max. dissipation. Exceeds MIL-R-11804C, Char. P.	S20 S25 S30	10 to 500 K 10 to 1.5 meg. 30 to 4.2 meg.	120° C. 40° C. ½ 1 1 2 2 4	±.03%/° C. from −55° C. to +235° C., ref. to 25° C.
R **Power.** Essentially non-inductive in high-frequency operations. Inherent noise level less than 0.1 microvolt per volt. Exceptional moisture resistance and overload capacity. Exceeds MIL-R-11804C.	R31 R33 R35 R37 R39	10 to 70 K 30 to 150 K 20 to 300 K 20 to 500 K 40 to 1 meg.	7 @ 40° C. 13 @ 40° C. 25 @ 40° C. 55 @ 40° C. 115 @ 40° C.	±.05%/° C. from −55° C. to +235° C., ref. to 25° C.

Why you save ounces and inches with Corning MIL resistors

Fuse a tin oxide coating to a piece of special glass. Spiral a helix in it. Attach leads. You have a unique resistor, smaller and lighter than other designs.

Why unique? Because the coating is an integral part of the glass base. It cannot come off or change its position during jarring or jouncing.

This simple, rugged, extremely reliable resistor will take all the environmental stresses created by the space age without changing its performance a whit.

The construction has exceptional low-noise and stable-temperature characteristics.

Above all, they are small. They are light.

The coupon will bring you complete technical data. Address: Corning Glass Works, 542 High Street, Bradford, Pa. For orders of 1000 or less, contact your distributor serviced by Erie Distributor Division.

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Company ....
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FXC makes computers
REMEMBER

Pictured above is a typical FXC coincident current memory stack; appearing, at left, one of the memory planes used in its assembly.

Time was when the problem of "remembering" placed practical limitations on the speed and capacity of computers. Not so, today, because of a FXC-developed component called the *coincident current memory stack.* Main reason for the outstanding success of this important advancement in magnetic ceramics rests in FXC's ability to meet the computer industry's requirements for ferroxcube cores for recording heads . . . pulse transformers . . . coincident current planes, stacks and similar *precisely engineered* products. Call FXC's Computer Engineering Dept. whenever you need help on a ferroxcube application.

Say
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need ferrite.

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50 East Bridge Street, Saugerties, New York

Section Meetings

(Continued from page 98A)

LANCASTER

"Some Aspects of Metallurgy in Semiconductors," Dr. F. D. Rosi, RCA. 12/9/59.

LAS CRUCES-WHITE SANDS

"Tracking Earth Satellites with Minitrack," P. A. Lantz, National Aeronautics and Space Administration. 2/9/60.

MEMPHIS

"Manufacture and Precision of Radio Crystals," Dr. Charles Ray, University of Tennessee. Election of officers. 1/30/60.

MONMOUTH

"Useful Energy from Controlled Fusion," Dr. M. A. Heald, Swarthmore College. 1/20/60.

NEW HAMPSHIRE

"Medical Electronics," A. M. Grass, Grass Instrument Co. 1/20/60.

NORTHERN VERMONT

"Submarine Electronics," Com. J. G. Gillmore, U. S. Naval Submarine School. 1/18/60.

ORANGE BELT

"Ionic Propulsion," Y. C. Lee, Aerojet General. 10/13/59.

"Quality Instrument Manufacturing," W. D. Meyers, Hewlett-Packard. Tour of California Polytechnic Electronic Laboratory Facilities. 11/10/59.

Tour of Kaiser Steel Fontana Plant and Plant Communications. Mr. Klussman, Kaiser Steel. 1/11/60.

PANAMA CITY

Business meeting. 2/10/60.

PASADENA

"Tunnel Diodes," Dr. C. J. Carter, Space Technology Labs. 1/8/60.

READING

"Industrial Communications," H. Campbell, GAI-Tronics Corp. Tour of Semi-conductor department of Western Electric Company plant in Laureldale. 11/18/59.

"Project Vanguard—A Step Into Space," H. F. Eckenroth, Martin Co. 12/11/59.

RICHLAND

"Elements of Reliability," A. L. Ruiz, G. E. Co. 11/4/59.

"The Precision Measurement of Extremely High Voltages," H. V. Larson, G. E. Co. 1/27/60.

SAN FERNANDO VALLEY

"You Take the Eyewitness, Give Me the Physical Evidence," Prof. L. V. Jones. Tour of Anheuser-Busch Brewery. 1/13/60.

SANTA ANA

"Earth Movements," John Healy, Seismological Lab. 1/12/60.

WESTCHESTER

"Application of Electronics to Photogrammetric Techniques in Modern Map Making," M. M. Thompson, U. S. Department of Interior. 1/20/60.

WESTERN NORTH CAROLINA

Two films: "Crystals—An Introduction" and "Brattain on Semiconductor Physics." 1/22/60.

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NEWS

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1959
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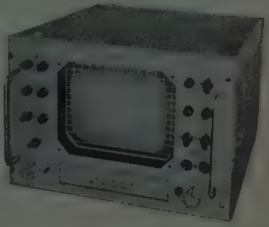
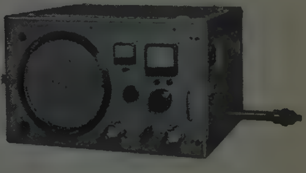


Displays two separate quantities such as impedance and gain as functions of frequency in the form of continuous curves. Frequency range: 500 kc to 400 mc. Instrument contains a sweep signal generator, precision variable attenuator, electronic switch, crystal marker generator and large screen oscilloscope which provides a complete precision measuring system.

Applications include laboratory and production testing of band-pass filters, limiters, all types of amplifiers, television receivers, attenuators, discriminators and coaxial cables.

Write for Bulletin SWOB.

DIAGRAPH



30 to 2400 mc. Plots, instantaneously and accurately by means of a light spot, complex impedances and admittances directly on Smith charts. Eliminates tedious measurements and involved calculations. Instrument can also be used as a phase meter over its frequency range.

Applications: impedance measurements on semi-conductors, antennas, filters, receivers, amplifiers.

Diagraph is available in three models: ZDU covering frequency range 30 mc to 300 mc and ZDU (420) from 30 to 420 mc; ZDD from 300 mc to 2400 mc. Overall accuracy is better than 3% for amplitude and 1.5° for phase angle.

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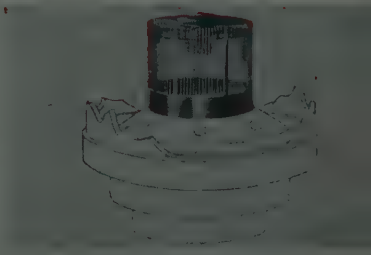
New RCA-7650 and -7651 Ceramic-Metal Beam Power Tubes promise dependable operation even when subjected to vibrational accelerations well over 2

When you're designing equipment for missile use, you can't take chances. You must be *sure* you're using the toughest tubes you can get. That's why the exceptional ruggedness of RCA-7650 and RCA-7651 is so important to you.

These two unique new ceramic-metal beam power tubes will actually withstand 20 g vibrational acceleration without adverse effect. In fact, both types have successfully undergone variable-frequency vibration tests (20-2000 cycles) with peak accelerations of twice that amount! Resistance of these tubes to vibration as well as to shock means dependable operation during the critical moments of launching and after. No wonder missile-equipment designers are so enthusiastic about these new tubes.

Both RCA-7650 and RCA-7651 feature a coaxial electrode structure which makes them adaptable to either coaxial-cylinder or parallel-line circuits. Both utilize RCA's exclusive grid-making technique for precision grid line-up and exceptional structural rigidity. Capsule data on these two forced-air-cooled types are shown in the adjacent chart. Technical bulletins on these types are available from RCA Commercial Engineering, Section D-35-Q, Harrison, N. J. For further information about these tubes and other RCA Ceramic-Metal Power Tubes, contact the RCA Field Representative at our office nearest you.



Type	Operation	Max Plate Volts	Max Plate input Watts	Max Plate Dissipation Watts	Useful CW Power Output Watts		Power Ga	
					at 400 Mc	at 1215 Mc	at 400 Mc	a 1215
RCA 7650	CW	2500	1250	600	800	450	14 db	7 d

Type	Operation	Max Plate Volts	Max DC Plate Amperes During Pulse with 10 μs duration and duty factor of 0.01	Max Plate Dissipation Watts	Useful Power Output at Peak of Pulse Watts at 1215 Mc	Pow Ga a 1215
RCA 7651	Screen and Plate Pulsed	8000*	9	600	39000	8 6
	Grid Pulsed	4000	9	600	20000	5

*Peak Positive

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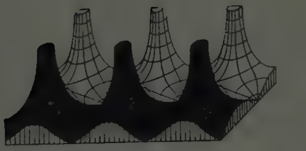
April, 1960
Vol. 48 No. 4
Proceedings of the IRE

Poles and Zeros



Space Electronics: "That which is characterized by extension in all directions," "of or pertaining to an electron or electrons," are Webster's definitions of space and electronics. Modern usage combines the two words, and in their modernity, as a single term, they have taken on a broad and very important scientific meaning. This, April, issue of the PROCEEDINGS reports a comprehensive and detailed definition of the term "space electronics" in its modern connotation.

For many months the public press has been filled with disquieting discussions of the pros and cons of the status of missiles and the entire "space program" of our nation. While discussion rages (and Generals disagree) it is fortunate that the scientist and the engineer move ahead with the task of finding the answers to many perplexing, interesting, and exciting problems.

The treatment of the subject to be found in this volume of the PROCEEDINGS resulted from an idea suggested by former Editor Ryder. The Editorial Board finding merit in the idea asked Conrad H. Hoeppner, Chief Scientist of Radiation Inc., to organize the issue. The wisdom of the Editorial Board's decision is demonstrated by the 53 impressive papers that he has assembled. Mr. Hoeppner is to be congratulated on the excellence of his work, and its results. The cover illustration for this issue was also Mr. Hoeppner's idea; he assisted and guided the preparation of the art work. He has added this chore to his other responsibilities as Editor, and as Chairman of the Professional Group on Space Electronics and Telemetry.

It is reassuring to learn of the extensive progress of research and investigation by the foremost workers in the area of space electronics. This issue marks another milestone in the IRE array of Special Issues.

Headquarters: IRE marks its continuing growth by the acquisition of the building at 984 Fifth Avenue; it is adjacent to and on the north side of its present building, 1 East 79th Street, in New York City. IRE now owns a group of three buildings on the northeast corner of Fifth Avenue and 79th Street. All originally belonged to the family of Irving S. Brokaw, famous 19th century clothier.

The Institute took occupancy of the initial building at 1 East 79th Street in 1946, and acquired the adjoining building at 5 East 79th Street in 1954. As is evident from the Table of IRE Growth, the necessity for additional space in the future becomes self evident. The purchase of 984 Fifth Avenue provides this space. There are no immediate plans for occupying the building, which is presently still in use as a private dwelling.

A study of the statistics in the Table yields some interesting results. At the end of 1953 Headquarters floor space was approximately 0.4 square feet per member. The time had arrived to provide additional space. Again, at the end of 1959 the same floor area per member has been reached and the new building acquired. Assuming a continuation of the efficient use of facilities the building just added should provide adequate floor space until the membership reaches 124,000. The acquisition of the new building thus assures that the rapid growth of IRE activities and services may continue unabated during the coming years. It is startling to discover, however, that if the recent rate of growth is maintained, the figure of 0.4 square feet per member may again be reached as early as 1965. A similar study of the data of the Table gives very roughly the figure of 1.8 square feet of Headquarters per page of publications and would indicate approximately 27,000 pages as the publication output in 1965. The Table would also predict a Headquarters staff of 260 in that year. The figures probably mean nothing in an Institute as dynamic as the IRE, but it's still fun to speculate.

TABLE OF IRE GROWTH

	12/31/45	12/31/53	12/31/59
IRE Membership	15,782	37,134	79,162
PG Members	0	21,797	87,027
IRE Sections	36	71	105
IRE Annual Expenses	$175,000	$1,300,000	$3,300,000
Publication Output (pages)	1,882	8,360	17,968
Headquarters Staff	32	85	170
Headquarters Floor Space (sq. ft.)	15,668	31,293	48,946

The building was purchased with IRE funds which were accumulated gradually over the years under a program of careful financial management. Our thanks go to our astute Treasurer, W. R. G. Baker, and to our Executive Secretary, George Bailey.

Better Paper: Beginning with the February issue, as an experiment, the same type paper has been used in the editorial section as had previously been used in the advertising section only. The principal difference is that the new paper has a glossy finish, providing a noticeable improvement in the reproduction of line drawings and halftones. In addition, the new paper is slightly heavier; thus the ink on one side of the sheet does not show through as much on the other side of the page. The new paper will tend to become more yellow with age, especially if exposed for long periods to sunlight, however, the amount of yellowing that is likely to occur to a bound volume in the library over a fifty year period will not be serious.

Convention Record: Now that the 1960 International Convention has taken its rightful place in IRE history, it is appropriate to issue a reminder that the Convention Record is no longer distributed free of charge. For your convenience a tear out order blank will be found on page 18A of this issue.—F.H., Jr.

J. A. Ratcliffe

Vice President, 1960

J. A. Ratcliffe (M'29–A'32-F'53) was born in Bacup, England, on December 12, 1902. He studied physics at Cambridge University and received the B.A. degree in 1924. He at once started research work, under E. V. Appleton, in the Cavendish Laboratory and assisted him, and M. A. F. Barnett, with their early experiments on radio waves reflected from the ionosphere. Soon afterwards Dr. Appleton left Cambridge and Mr. Ratcliffe continued his ionospheric researches, first alone, and later with a group of research students.

During World War II, he worked with the Telecommunications Research Establishment (T.R.E.) which was concerned with the design and development of Airborne Radar. When the war was over he returned to his radio research at the Cavendish Laboratory. He was joined there by Martin Ryle, who very soon established his own research group of Radio Astronomers.

Mr. Ratcliffe is at present a Fellow of Sidney Sussex College and a Reader in Physics in the University of Cambridge. On October 1, 1960, he will become Director of Radio Research in the Department of Scientific and Industrial Research (D.S.I.R.). He will leave the University and will assume control of the Radio Research Station, Slough, which has had its terms of reference extended to include exploration of the upper atmosphere and outer space by radio and nonradio methods.

His research, and that of his research group, has been concerned almost entirely with the physics of the ionosphere. He has been particularly interested in the magnetoionic theory (on which he has written a book), ionospheric cross modulation, the diffractive reflection of radio waves from an irregular ionosphere, horizontal movements of irregularities in the ionosphere, the computation of electron distributions in the ionosphere, and theories of the formation of the *F* layers.

Mr. Ratcliffe was made a Fellow of the Royal Society in 1951, and a Commander of the Order of the British Empire (C.B.E.) in 1959. He is at present President of the Physical Society and Vice-Chairman of the Electronics and Communications Section of the Institution of Electrical Engineers. He is Chairman of the British National Committee for Scientific Radio which is responsible for arranging the XIIIth General Assembly of URSI in London in September 1960.

Space Electronics*

CONRAD H. HOEPPNER†, FELLOW, IRE

In lieu of the usual "Scanning the Issue" page, the organizer of this special issue presents below an introduction to the subject of this issue, followed by a brief outline of the contents.—*The Editor.*

THE SPACE AGE

A NEW AGE is beginning in the history of mankind—an age in which man is no longer confined to his planet, but is free to travel through the infinite vastness of space. At first he will find this new freedom dangerous and the environment hostile, but as his knowledge increases and he attains mastery of space, the freedom to travel will become safe and the environment friendly. In the new age are limitless possibilities for adventure, exploration, business, pleasure—and war.

The capability to explore space has developed largely through the military efforts of the United States and Russia. While the art of rocketry is old, it has only recently been exploited to the point of space travel. In World War II, German scientists demonstrated a weapon of potentially great effectiveness in the rocket which we know as the V-2. Both the United States and Russia have employed these scientists on their respective rocketry programs and have added their own manpower and other resources to make the programs effective. Peaceful and scientific uses of space probes and satellites have developed through the International Geophysical Year plans and then through the establishment in the United States of the National Aeronautics and Space Administration. The struggle now is to obtain more thrust per pound from chemical fuels and to augment or replace chemical fuels with nuclear and electronic means of propulsion. As these efforts meet more and more success, larger payloads of equipment and men may be taken deeper into space—and returned.

While initial efforts are to obtain more thrust per pound, such a result must be obtained in a manner which will insure reliability and safety. The many hazards of space, among which are dangerous radiation and collision with other objects traveling at high speed, must be surmounted. In order to sustain life, a closed régenerative cycle for producing food, water and oxygen must be perfected and it must function, even in the event that initial supplies are inadvertently destroyed. After survival of a journey in a space vehicle to the moon or another planet, suitable places in which to live must be found, developed and maintained. These places are likely to be underground caves instead of the glamorous "bubbles" so often portrayed by fiction writers. Other unpleasantness and hardships of living on extraterrestrial bodies will develop, but human ingenuity will triumph and eventual success in the colonization of other bodies in space is certain.

Man is a self-sufficient species on earth. There are many portions of his own planet not adequately explored and certainly not colonized. It was not really necessary that he explore and colonize the Americas. He needn't make expeditions to the Poles in order to live amply. It is intrinsic in the nature of man, however, that he is not content with past accomplishment. He must continue to strive for the unattainable and he will not rest even when he has possessed every secret of the universe.

* * *

We *will* become space travelers!

We will establish adequate reserves of energy (enough thrust per pound) for travel through space and with the necessary reliability and safety to transport human beings. We will develop protective measures against radiation, collision and other hazards. We will produce suitable regenerative cycles for food, water and oxygen to sustain life reliably. We will provide rapid transportation to man-made satellites, natural satellites, and other planets, and we will develop adequate and even luxurious living accommodations to colonize other extraterrestrial bodies. These accomplishments, however, will demand still greater efforts from the men and nations of the earth. Present efforts fall short *by orders of magnitude.* In the United States, for example, space efforts, measured in dollars, are approximately one quarter of the automobile production. These efforts must increase over one-hundred fold before space travel is common!

WHY EXPLORE SPACE?

Certainly, it has been the military advantage of intercontinental ballistic missiles which has spurred the development of vehicles capable of the exploration of space. Space vehicles, too, have military values both for surveillance and as weapons. Participation in the International Geophysical Year also has led to the uses of rockets and satellites for obtaining scientific data. The physical knowledge of the space surrounding the earth and of the earth itself was increased. Political prestige

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* Original manuscript received by the IRE, January 11, 1959.

† Radiation Inc., Melbourne, Fla.

has further been a stimulus to the development of space vehicles. The Russian successes have caused the United States to increase its rather minimal space exploration program. Then, of course, there are new territories to "homestead" and new sources of wealth in minerals and raw materials to be discovered. Whether the reasons for space exploration be as listed above, or whether they be for satisfying man's curiosity and his desire to advance, space exploration has begun. It now remains for us to examine how it shall continue.

PLANS FOR SPACE EXPLORATION

Initial exploration of space was with unmanned probes, *viz.:* high-altitude rockets, satellites, and deep-space exploration vehicles. These were supplemented by the terrestrial-based measurements of optical astronomy, radio astronomy, and cosmic particles. These methods will continue, but unmanned space probes will be followed closely by manned vehicles. Such vehicles will travel first around the earth, then to satellite space stations, to the moon, later to other planets, to their moons, and to the planetoids. Combinations of manned and unmanned space probes will also be useful. One plan for visiting the moon calls for the landing of several unmanned alternate vehicles on the moon to insure the safe return of passengers when the manned vehicle finally arrives.

Satellite space stations are a likely development and will serve as refueling stations for further travel, as well as for terrestrial communications. These stations may be both near and far from the earth. A moon base would certainly be a desirable launching point for further space probes and is an ideal base for radio and optical astronomy.

Planetoid bases with their lesser gravities may be extremely valuable for space exploration. They have sufficient volume to provide living quarters and raw materials for construction. Their masses may be used as protection from the radiation hazards as well as particle impacts. Their greatly eccentric but highly predictable orbits make them ideal for further travel to planets.

As to the planets, both Mars and Venus present the possibility of colonization. Mercury may also be habitable. Other planets, because of their high gravity or remote locations, will require much more in the way of propulsive energy and protection for landings or exploration and may be quite unsatisfactory for habitation. The moons of Jupiter and Saturn are more likely targets for exploration and colonization than those planets themselves.

In the exploration of space, the role of electronics is most vital. Electronic measurement and transmission of data are the forefronts of our exploration effort. Recording and analyzing equipment are also largely electronic. Many new equipments and systems are needed for measurements, telemetry, recording and processing. The techniques developed in the past for rocketry must be extended and improved. The need to reduce size, weight and power requirements of the equipment is paramount in the light of our present feeble rocketry capability.

COMMERCIALIZING SPACE

The first uses of satellites probably will be for surveillance. The photographs of the back side of the moon taken by Russian scientists are a typical example. Optical surveillance as well as infrared, radio, and radar surveillance of the earth and other bodies are in demand for military applications, for surveying and for predicting and possibly controlling the weather. A second use of satellites is likely to be the relaying of communications, permitting a vast increase in the number of intra-earth message channels. Multiple station coverage may be controlled for relayed communication as well as for recorded and delayed communication. Satellite usefulness will undoubtedly be extended to include radio and television rebroadcasting with worldwide coverage. This extensive scope of broadcasting will bring about interesting and revolutionary changes in the methods of handling and regulating these facilities. The satellites will also widen our research horizons, as will colonization of the moon and other planets. Astronomical data may be better and more easily obtained on the moon than it is on earth. The ease of providing a vacuum for manufacturing and experimental processes is an obvious advantage. Thermal insulation and the obtaining of low temperatures are radically simplified from conditions on the earth. As a result, it is likely that the space experimental laboratory will differ greatly from its earthly counterpart. Also possible are new manufacturing processes that may produce improved life on earth as well as in space.

FUTURE PROBLEMS

The space age has barely begun, yet almost insurmountable problems have developed. More and more serious problems may be expected in the future.

At first will be the scientific problems. Physical, chemical, biological, and medical laws and limitations will be widened and stretched, possibly beyond our present expectations. On the electronic side, radio-frequency propagation anomalies which are not now suspected will undoubtedly be discovered. The new space environment will cause design and operating changes in practically every branch of electronics.

However, these technical problems may well be obscured by man-made problems unless we formulate regulations and legislation designed to govern the activities of the astronaut. There must be laws common to the many participating countries and additional laws to govern the new encounters in space. International agreements which have been heretofore difficult, if not impossible, to make are necessary to prevent chaos. The social problem of people living in close quarters for long periods of time with mental strains, boredoms, etc., racial or national prejudices, and the possibility of new nations and new governments must be considered. A careful and considered program of space law is a necessity.

In the area of electronics there will be the frequency allocations to prevent interference, and, of course, in radio and television broadcasting, who is to regulate the quality of the programs? As a matter of fact, will the Federal Communications Commission of the United States permit unattended satellite repeater broadcasting stations? What will happen to the "spheres of local interest" which the FCC has fought valiantly over the years to preserve? Installing unattended repeater radio and television broadcast stations all over the United States would be far more economic at the present time than using a repeater satellite. Licenses, however, are not presently granted for this purpose. Also, what is to prevent the unauthorized use of a repeater satellite, a use which may interfere with its normal purpose? It is accessible from every point over half the globe.

Applications of electronics which were not even conceived a year ago are already realities. It is these problems and scientific aspects of space exploration with which the current issue of the PROCEEDINGS OF THE IRE is concerned.

THE APRIL, 1960, SPACE ELECTRONICS ISSUE

Many of the equipments and techniques used in space research and satellite programs are adaptations of those employed in earlier rocket experiments, but on the other hand, some are new. In this latter category is electronic propulsion. Early in the issue this subject is introduced, its merits discussed, and comparisons are made with existing chemical propulsion systems.

In the categories of navigation and guidance, a brief introduction to the solar system is presented and methods of navigation and guidance for interplanetary travel are given. In addition, the use of satellites for navigation of ships, planes and other vehicles on the earth's surface is treated. A theoretical article is presented in which the author develops the thesis that present inertial guidance systems may improve by as much as two orders of magnitude.

Communication from point to point on the earth, from earth to satellite, to interplanetary traveler, and to other planets is considered. In this very broad field of communications, only a few representative and diverse subjects could be included. Tracking and surveillance papers are presented which illustrate a small portion of this field.

Telemetry in space is examined from a number of aspects. As usual, the paramount problem is that of transmitting as much data as possible for as great a distance as possible with a minimum of equipment.

A representative group of papers representing instruments and measurements has been chosen, ranging from those necessary to maintain and direct the vehicle to those used to discover the physical conditions in outer space.

While space electronics is introduced to readers in this issue of the PROCEEDINGS OF THE IRE, the TRANSACTIONS of the Professional Group on Space Electronics and Telemetry will continue to present these subjects to its members. Six basic divisions of the PGSET are active in collecting and disseminating information in the following categories:

1) Telemetry and Remote Control.
2) Space Instrumentation.
3) Space Communications.
4) Space Navigation.
5) Electronic Propulsion in Space.
6) The Effect of Space Environment on Equipment and Living Things, Including Humans.

If any of these subjects is of interest, a cordial invitation is extended to our readers to continue that interest with the PGSET and its TRANSACTIONS.

The NASA Space Sciences Program*

Summary—**This report, prepared in April, 1959, summarizes the objectives of the space sciences program of the U. S. National Aeronautics and Space Administration and the immediate and long-range plans for carrying the program out. The report is divided into several broad areas: atmospheres, ionospheres, energetic particles, electric and magnetic fields, gravitational fields, astronomy, and biosciences. Included in the discussion is a review of present knowledge and existing problems in each of the above areas.**

THE U. S. National Aeronautics and Space Administration is developing a national space sciences program on as broad a basis as possible. In the planning and programming, advantage is being taken of the advice of many specialists and experts in the scientific community. In the conduct of satellite and space probe experiments, broad participation of the scientific community and industry, along with government, is planned, and steps are being taken to secure such participation. The developing program has captured and increased the momentum in space research developed during the International Geophysical Year.

The present status of the national space science program is described in the following sections. For convenience, the subject of research in space has been divided into several broad areas: Atmospheres, Ionospheres, Energetic Particles, Electric and Magnetic Fields, Gravitational Fields, Astronomy, and Biosciences. For each area, the principal scientific objectives are stated, followed by a review of present knowledge and existing problems to be solved. Each section closes with a statement of the long-range program and of the immediate program now under way.

The subdivision used is one of convenience only. Actually, the various areas are related in many complex ways. Investigations of the various interrelationships are a very important part of the search for a better understanding of many puzzling phenomena.

It is recognized that the national space sciences program is subject to continuing review, amplification, and modification as work progresses and scientific data and results are obtained from currently-orbiting space vehicles. Moreover, desired scientific experiments must continuously be matched with the present and future capabilities and availabilities of vehicles, supporting technology, and facilities.

ATMOSPHERES

Objectives

To determine and understand the origins, evolutions, natures, spatial distributions, and dynamical behaviors of the atmospheres of the earth, moon, sun, and planets; and their relations to the medium of interplanetary space. To investigate atmospheric phenomena associated with interactions between photons, energetic particles, fields, and matter. To understand the relations between the earth's upper atmosphere and its surface meteorology. To evaluate atmospheric effects on instrumented and manned space flight.

Present Knowledge

The Earth's Atmosphere: The structure, *i.e.*, temperature, pressure, and density, of the earth's upper atmosphere is partially known from indirect observations and from rocket soundings at New Mexico, Fort Churchill, Canada, in the Arctic, and on the equator. Up to about 110 km altitude, density and pressure both fall off by a factor of roughly 10 for every 10-mile increase in height. At higher altitudes, the rate of decline in density and pressure is markedly slower. Starting from the ground, temperature falls steadily with height until at about 12 km (in temperate latitudes) the temperature reaches a minimum of about −55°C. With increasing altitude, the temperature then rises again to a maximum of about 0°C at between 50 and 60 km. After that, there is a sharp decline in temperature to probably less than −75°C at between 80 and 85 km height. Throughout the E and F regions of the ionosphere, kinetic temperature rises to 1000°C or more, and perhaps to above 2000°C. The structure parameters vary with both time and geographic position. Pressure, for example, varies by a factor of at least two at 100 km and by a factor of six or more at 200 km. Indications from rocket measurements are that variations by orders of magnitude in pressure and density may be expected at F-region altitudes and higher.

The composition of the atmosphere is essentially the same from sea level up to the E region at 100 km. Small amounts of ozone are created by solar ultraviolet radiation, the maximum concentration being found at between 25 and 30 km in the daytime. In the E region and above, molecular oxygen dissociates to atomic form. Rocket observations at Fort Churchill have shown a marked diffusive separation of argon relative to nitrogen starting at altitudes between 100 and 120 km and increasing into the lower F region.

Rocket wind measurements and indirect ground-based observations have indicated fast-moving winds at all levels of the upper atmosphere. Typical speeds are 20 to 50 meters per second between 30 and 120 km, while speeds higher than 100 meters per second have

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* Original manuscript received by the IRE, December 2, 1959. Prepared by the Office of the Asst. Dir. for Space Sciences, Natl. Aeronautics and Space Admin., Washington 25, D. C., April 16, 1959.

been measured. In general, the winds are from the West and strong in winter and from the East and weaker in summer. Marked wind shears have been observed in the *E* region of the ionosphere.

The Sun's Atmosphere: The sun's atmosphere may extend with decreasing density throughout the solar system. At a distance of about 20 solar radii (nearly 0.2 Astronomical Units) the atmosphere is still directly observable by its blurring effects on the radio stars seen through it. Near the earth's orbit, it is estimated that this atmosphere contains several hundred protons per cubic centimeter and has an electron temperature of 250,000°K.

The Moon's Atmosphere: The moon lacks any appreciable atmosphere, although a highly rarefied ionosphere has been indicated by radio astronomical observations.

Planetary Atmospheres: Mercury has no observable atmosphere, and Pluto is too faint for an atmosphere to be detected through the terrestrial atmosphere. Atmospheres have been observed on each of the other planets and on one of Saturn's satellites.

Venus is covered by an extensive atmosphere of which CO_2 is the primary constituent. H_2O has not been identified there although it may be present below the opaque cloud cover which envelops the entire planet. The temperature of the top of the cloud layer has been measured with a thermopile as about 0°C. Radio-astronomy measurements indicate a temperature of nearly 300°C lower in the atmosphere.

The Martian atmosphere is tenuous compared to the earth's, but H_2O and dust clouds have been observed there. The surface pressure has been estimated at approximately eight per cent of the earth's sea-level atmospheric pressure.

The atmospheres of the major planets (Jupiter, Saturn, Uranus, and Neptune) contain large amounts of H_2 and, presumably, He. Their spectra are dominated by bands of CH_4 and NH_3. The atmospheres of Jupiter and Saturn have a marked banded structure and exhibit differential rotation as does the solar photosphere. In addition, Jupiter shows such puzzling features as the famous red spot, white spots, and lightning-like radio emissions.

Existing Problems

The Earth's Atmosphere: Major problems connected with the earth's atmosphere are exemplified by the following questions, the answers to which are known only partially or not at all.

1) What are the primary sources of energy affecting the high atmosphere, and at what altitude levels are these energies deposited in the atmosphere?

2) What is the average structure of the atmosphere above 80 km over the entire globe; and what are the variations in the structure parameter with respect to season, time of day, geographic location, and solar activity?

3) What is the detailed composition of the atmosphere above 100 km? At what altitude, if any, do the light gases H, H_2, and He predominate over O, N_2, and N?

4) What are the detailed dynamics of the high atmosphere, including the general circulation?

5) What is the relation of the Great Radiation Belt to the heating of the upper atmosphere, and what other sources and sinks of energy are there?

6) What is the energy budget of the low atmosphere and what is its relation to cloud cover and weather systems?

7) What are the relations between the earth's surface meteorology, its upper atmosphere, and solar activity?

8) What is the origin and history of the earth's atmosphere?

The Sun's Atmosphere: Similarly, these questions relate to the sun's atmosphere.

1) What is the composition of the solar atmosphere, particularly at planetary distances?

2) What are the density and temperature of the sun's atmosphere as a function of distance from the sun?

3) How does the sun's atmosphere interact with those of the earth and the other planets?

The Moon's Atmosphere: In this context, two major questions about the moon are raised.

1) Has the moon a trace atmosphere and, if so, what is its composition?

2) If the moon does have a trace atmosphere what are its origins, dynamics, and entrapment near the moon's surface?

Planetary Atmospheres: Finally, the following questions arise concerning the atmospheres of the other planets.

1) What are the major constituents of the planetary atmospheres?

2) What are the surface pressures and densities?

3) What is the altitude variation of their atmospheric structures?

4) What are the atmospheric components that might support earth-type life?

5) Do their atmospheres contain oxidizers available for combustion?

6) Does Venus's atmosphere contain H_2O?

7) Of what does the cloud cover of Venus consist?

8) What causes the apparent changes in transparency of the Martian atmosphere, so-called "blue clearing"?

9) What is the nature and cause of the famous red spot on Jupiter?

10) What is the nature and origin of the rings about Saturn?

11) What are the important dynamic features of these atmospheres?

12) Has Pluto an atmosphere?

Program

Long-Range: Over-all, long-term plans for achieving the above objectives include: 1) instrumented satellite stations around other planets; 2) rocket probes deep into the atmospheres of other planets including soft landings onto the surface with automatic and, eventually, manned recording stations; 3) probes deep into the solar atmosphere; 4) special probes for measuring the density and nature of gas and dust particles in interplanetary space and within comets; and 5) extensive theoretical studies to understand the basic natural phenomena taking place within the atmospheres.

Immediate: Short-range plans include extensive and intensive studies of the structure and composition of the earth's atmosphere from 80 km to several hundred kilometers by direct measurements with sounding rockets and with satellites. Diurnal, latitudinal, and temporal variations in these parameters will be determined and will be correlated with the energy and momentum balances in the earth's upper atmosphere. Models of the earth's atmosphere will be formulated for 1) providing basic data needed in understanding ionospheric, auroral, and other phenomena, 2) providing guidance in the study of the atmospheres of other planets, and 3) providing essential information in the engineering design of the man-in-space launching, guidance, and recovery systems.

Short-range plans for studies up to 80 km include a large number of synoptic rocket flights and several cloud-cover satellites to establish the relationship between surface meteorology and the structure and dynamics of the upper atmosphere. These data are vitally needed to improve weather forecasts and to permit eventual control of certain aspects of the weather.

Immediate steps in the atmospheres research program are: 1) initiation of extensive instrumentation and systems development contracts for measuring atmospheric structure and composition at gas densities of from 1 to 10^{-13} atmospheres; 2) adaptation of currently available atmospheric-structure instruments for an unmanned earth satellite; 3) continuation and improvement of IGY rocket soundings to determine the vital factors in the large variations noted in the IGY atmospheric-structure flights; 4) the development of an inexpensive synoptic meteorological rocket system to permit rapid collection and analysis of data up to 80 km; 5) rocket test flights for proposed satellite instrumentation systems; 6) launching of the remaining Vanguard meteorological satellite to study radiation balance in the earth's atmosphere; 7) completion of the test phases and flight phase of the ABMA radiation balance satellite; and 8) the launching of a 12-foot inflatable sphere to obtain satellite-drag data from which atmospheric densities can be computed.

IONOSPHERES

Objectives

To determine and understand the sources, natures, spatial distributions, and dynamical behaviors of the ionized regions of the solar system, including the ionospheres of the earth, moon, and planets. To investigate ionospheric phenomena resulting from interactions between photons, particles, ions, and magnetic, electrostatic, and electromagnetic fields. To understand relations between solar activity and the terrestrial and other planetary ionospheres, magnetic fields, and upper-atmospheric current systems. To evaluate ionospheric effects on instrumented and manned space flight, including communications.

Present Knowledge

The terrestrial ionosphere, which extends from an altitude of 60 km into interplanetary space, is formed by the ionization of neutral particles, primarily by radiation emitted from the sun. The density of ionization is not uniform. The altitude regions of maximum ion density during daytime are designated D, E, F_1, and F_2. Additional sporadic ionization can occur at various levels, particularly in the E region. Normal D-region ionization extends roughly from 60 to 85 km, with maximum values no greater than a few thousand electrons per cubic centimeter. Average altitudes at which maximum densities in the E, F_1, and F_2 regions occur are 105, 150, and 300 km, corresponding to concentrations of 1.5, 2.5, and 20×10^5 el/cm^3. These values are for a summer day in New Mexico during the period of maximum solar activity. Gradients in the lower E and F_1 regions are roughly 10^4 el/cm^3 per km. Very recent results indicate that the concentration above the F_2 region experiences a very gradual decrease to a value of 10^3 el/cm^3 at a height of 10,000 km.

At night the density of ionization in the D, E, and F_1 regions decreases by at least an order of magnitude. The concentration decreases by less than this factor in the F_2 region, with the result that the nighttime ionosphere shows a single F region predominant with comparatively little ionization below 170 km.

The E-region altitude is relatively insensitive to latitude. However, the F-region height is greatest over the equator and decreases by 100 km at temperate latitudes. Ionospheric irregularities are strongly dependent on geomagnetic latitude, with extremely irregular conditions occurring in the auroral zones. The electron density varies strongly with the solar sun-spot cycle, by as much as one order of magnitude in the F_2 region. The major ionized constituents of the E and F regions are nitric oxide and atomic oxygen, respectively.

The ionosphere makes possible communication over large distances by reflection of frequencies below the critical frequency. Typical values of these critical frequencies are 3 to 4 mc for the E region and 5 to 15 mc for the F region. It should be emphasized that since ionospheric structure is dependent upon diurnal, seasonal, latitude, and solar effects, so are its propagation characteristics. In the lower ionosphere, such phenomena as photoemission, diffusion, and the rectifying action by antennas result in charging space vehicles with potentials up to 20 volts.

Very little is known about the ionospheres of other planets. There are indications of thunderstorms on Jupiter and possibly Saturn and of auroras on Venus.

Existing Problems

Although ionospheric structure is being studied by vertical-incidence radio soundings from surface stations, this method gives results only up to the F-region maximum and provides only integrated effects. Past experience has shown that rocket and satellite measurements of electron density vs actual height are needed to interpret surface ionosphere sounding records correctly, and thus permit their use on a practical synoptic basis.

Since rocket measurements have not been made in sufficient number above the F_2 region and since they have intercepted only an extremely small portion of the ionosphere for small time intervals, the most important unknowns are the details of ionospheric structure, particularly above the F_2 region, as a function of time, season, latitude, and solar and cosmic activity. Collision frequencies and the ionic composition (particularly at low mass numbers) require special investigation.

Present knowledge of ionospheric-propagation characteristics requires extension. Needed are: explanations for anomalous fading effects; polar diagrams of upcoming and downcoming waves and of waves along the magnetoionic ducts; experimental verification of the theories of whistler and Z-mode propagation; the frequencies at which heavy ions influence propagation (hydromagnetic waves); and investigation of the probable existence of very low frequency windows useful for interplanetary communications. Definition of ionospheric structure and propagation characteristics is important to such problems as the calculation of tracking errors in AICBM radars and the choice of communications frequencies.

The interaction between charged particles in the ionosphere with space vehicles is one of the least understood phenomena of space flight. Measurements of vehicle charge, particularly in the Great Radiation Belt and in regions where photoemission effects can become serious (exosphere), are needed to compute possible electrical drag and missile-trajectory errors.

Propagation characteristics of the exosphere and the intensity and nature of cosmic noise need investigation, particularly in the frequency range which is most affected by the ionosphere. These data are important to radio-astronomy observations from space vehicles.

Since so little is known about other planetary ionospheres, discovery and survey-type investigations of them are needed first. Then their structure and propagation characteristics also will need to be studied.

Program

Long-Range: The over-all program will exploit current techniques for determining the terrestrial ionospheric structure, its propagation characteristics, and its influence on space flight, by observation from below, within, and above. New techniques for evaluating the least known parameters will be developed. All of the applicable methods will then be used for the study of other planetary ionospheres. Eventually, propagation sounding stations may be established on the surface of the moon. All the data will then be applied to understand the interrelations between solar activity, magnetic fields, the aurora, the Great Radiation Belt, and other phenomena.

Immediate: The immediate program is concerned with obtaining electron-density profiles at altitudes above the F_2 layer by propagation experiments in sounding rockets. Then, latitude and temporal variations of this parameter will be obtained by use of a satellite beacon. Topside sounders in satellites will be used for synoptic studies of electron density in the outer ionosphere. This technique promises less ambiguity than that obtained from satellite beacons. Present knowledge of electromagnetic propagation will be extended by the inclusion of VLF receivers in polar-orbiting satellites and space probes. Ion-spectrum studies will be extended to lower mass numbers and higher altitudes by the inclusion of RF mass spectrometers in space probes and satellites. Direct measurements using devices such as antenna probes, ion probes, and electric field meters will be made in rockets and satellites, to define more precisely ionospheric structure and to study the interaction between the ionosphere and space vehicles.

ENERGETIC PARTICLES

Objectives

To determine and understand the origins, natures, motions, spatial distributions, and temporal variations of particles having energies appreciably greater than thermal. To understand their relation to the origin of the universe. To understand interactions between such particles, fields, photons, and matter. To evaluate possible hazards to life and other effects of energetic particles and photons on instrumented and manned space exploration.

Present Knowledge

Cosmic rays are an important constituent of the energetic charged particles which are known to be present in space. It is now certain that cosmic rays consist primarily of high-energy hydrogen nuclei, plus a small percentage of energetic nuclei of heavier atoms. The heavier cosmic rays are largely energetic helium nuclei but include masses up to that of iron. It has been determined that the cosmic-ray energies cover at least the range from a few billion to a billion billion electron volts. Over this energy range the spectrum is such that the cosmic-ray intensity falls rapidly with increasing energy. Because of this spread of individual particle energies, and because of the magnetic field of the earth, fewer cosmic rays strike the earth's upper atmosphere at the equator than at higher latitudes.

On striking the atmosphere, the cosmic rays produce many secondary particles including electrons, many types of mesons, and fragments of atomic nuclei. Most of these secondaries are absorbed by the atmosphere, but a few reach the earth and some scatter back out of the atmosphere. The intensity of these secondary particles just above the atmosphere is comparable with the cosmic-ray intensity but decreases as the distance above the atmosphere increases. The cosmic ray intensity is not a constant but varies with the sun spot cycle and with magnetic storms and solar flares.

Another group of energetic particles in space comprises the particles which produce auroras. These particle streams contain electrons having energies over at least the range from a few hundred to a few thousand electron volts. Also present are hydrogen nuclei having energies over about the same range as the electrons. On striking the atmosphere, these particles ionize and excite the air atoms producing the light we see as auroras.

A third class of energetic particles known to exist in space comprises the Great Radiation Belt. These particles are in a belt surrounding the earth at latitudes less than 70° and at altitudes between about 600 and 30,000 miles. The intensity of the radiation in the belt is thousands of times that of the cosmic rays. It seems clear that the radiation belt is the immediate source of the auroral particles.

Existing Problems

Among cosmic-ray problems, important phenomena to investigate are: The sources of the primary cosmic rays, the processes by which the particles obtain their high energies, their distribution in space, their variations in intensity and composition with time, their energy spectrum particularly at lower energies, their interrelations with the auroral and radiation belt particles, their effects on both animate and inanimate objects, and their detailed composition including the possible presence of antimatter.

In the field of auroral particles, the following are unknown: The origin and acceleration mechanisms of these particles, their presence or absence in space and near other planets, their rapid variations in intensity and spatial distribution, their detailed composition, and their relations with cosmic rays, radiation belt particles, magnetic fields, the sun, the ionosphere, and with upper-atmosphere processes.

Among the unknowns associated with the Great Radiation Belt particles are: The composition and energy of the particles, their origins, their lifetime, their spatial extent, their possible variations in composition and intensity, their effects on animate and inanimate objects, and their relations to cosmic rays, auroral particles, magnetic fields, the sun, and upper-atmosphere processes.

Program

Long-Range: In situ measurements using deep space probes will be made from the close proximity of the sun to the limits of the solar system. Extensive measurements in the vicinity of the planets, especially the earth, will be made to determine the interaction of the energetic particles with the atmospheres and fields of these bodies. These measurements will require satellite orbits around the earth, the moon, and other planets. The establishment of an observatory on the surface of the moon or on some other planet might be necessary, depending on the data previously acquired by artificial satellites.

Immediate: In the near future, the measurements of energetic particles will be pursued with satellites and rockets in the vicinity of the earth and with interplanetary probes. These measurements will be aimed at determining the interactions of these particles with the earth's atmosphere and field, their interactions with interplanetary fields, the types and energies of these particles, their spatial distribution, and the origin of the energetic particles.

The immediate program includes, specifically, measurements of 1) the cosmic-ray intensity in interplanetary space; 2) time and latitude cosmic-ray intensity variations; 3) the composition and spatial extent of the Great Radiation Belt; 4) the cosmic-ray energy and charge spectrum; and 5) the nature of the particles producing auroras.

ELECTRIC AND MAGNETIC FIELDS

Objectives

To determine and understand the origins, natures, methods of propagation, spatial distributions, and temporal variations of magnetic and electric fields throughout the universe. To understand interactions between these fields and matter in space, and the influence of existing fields on solar and planetary atmospheres. To use these fields in the investigation of the internal constitution of astronomical bodies. To evaluate their effects and interactions on instrumented and manned space exploration.

Present Knowledge

The earth's main magnetic field is approximately that of a dipole located at the earth's center but tilted at an angle of 11 degrees with respect to the earth's axis. The strength, orientation, and uniformity of the main field varies slowly with time. This is called *secular variation*. The field at the earth's surface is also geographically irregular because of differences in the geology of the earth's crust. These irregularities are called *anomalies*. On a time scale of months rather than years, the above features appear constant at a given location. However, using a time scale of days, hours, or minutes, a sensitive instrument shows that the field is practically never constant at any location on the earth's surface. These variations are usually classified as: *diurnal*—the variations that occur in 24-hour cycles and appear similar from day to day; *magnetic disturbances*—large variations that are not a daily event, in which the field changes are often highly irregular for periods of several or more hours over either small or large areas of the earth's surface; *magnetic storms*—large variations which occur more or less simultaneously over the entire surface of the earth; *sudden commencements or impulses*—abrupt changes in the magnetic field which sometimes precede magnetic storms; *pulsations*—sinusoidal oscillations of the magnetic field, with wave periods ranging from fractions of a second to five minutes and frequently persisting for one or more hours and sometimes as long as a half a day. The magnitude and frequency of occurrence of these variations are functions of geomagnetic latitude. The largest diurnal variation is observed at the magnetic equator. The largest disturbance and storm variations occur in the auroral zones in association with visible auroras. Sudden commencements and impulses tend to be world-wide. Pulsations are most prominent in high latitudes and reach their greatest magnitude in the auroral zones.

From the study of surface measurements and from theory, it is believed that these variations are caused primarily by electric currents in and above the ionosphere. The currents are attributed to the dynamo action of tides and winds in the ionized gas and to the influx of particles from the sun. Recent thinking on hydromagnetic phenomena has led some theorists to believe that interactions between the outer portions of the earth's field and the motion of solar gas clouds also contribute to the variations. It has not, however, been possible to prove any theory or establish an understanding of the phenomena solely from surface measurements. Here lies the importance of direct measurements from space vehicles.

Sensitive magnetometers have been successfully flown into and through the *E* region. These flights demonstrate the existence of electric currents in the lower ionosphere at the magnetic equator and in the auroral zones and showed the expected decrease in field intensity with altitude. More than anything else, they show the need for more measurements.

The magnetic fields of the sun and other stars have been studied intensively, but it cannot be said that there is concrete knowledge of their strength and form. Indirect measurements and theories permit a wide range of possibilities. Galactic field intensities have been estimated from theoretical considerations and the study of light polarization; values between 10^{-3} and 10^{-6} gauss have resulted. Theoretical studies of planetary fields (other than the earth's) apparently do not exist, and methods of indirect measurement have not been found.

An electric field exists between the earth and its ionosphere, corresponding to an average potential difference of 290,000 volts. There is no direct knowledge of the existence of magnitude of electric fields outside the ionosphere. Theories of the origins and accelerations of auroral particles and cosmic rays allow a wide range of possibilities for interplanetary electric-field values.

Existing Problems

The unknowns of most urgent interest are usually those which are most apparent. The most apparent unknowns in magnetic field phenomena are the sources of field variations and the relationships between these variations and other solar-terrestrial phenomena. It is known that some variation is due to ionospheric currents; but the existence, location, and strength of possible electric currents above the ionosphere are not known even though their existence appears probable. Strictly speaking, the cause of ionospheric currents is not known; and even less is known about probable causes for currents above the ionosphere.

The form of the earth's main field is known, at least approximately, close to the earth; but at distances of 5 to 15 earth radii, where the field is most effective in guiding solar particles, radical departures from prediction can be expected. The field may expand and contract in response to motions of solar and interplanetary gas clouds. These distortions may in turn propagate as hydromagnetic waves down to the ionosphere and even to the earth's surface. The possibilities can only be considered speculative until field measurements are made in the outer reaches of the earth's field.

At some distance, the earth's field will blend with interplanetary and/or interstellar fields. Estimates of interplanetary fields range from 10^{-3} to 10^{-6} gauss. The strength and orientation of these fields is especially important to the understanding of how solar particles reach the earth.

Theories concerning magnetic fields on other planets are practically nonexistent. One reason for this scientific void is probably the lack of understanding of why the earth has a magnetic field. It is generally thought that the earth's main field is due to dynamo currents in the liquid core, but the form and cause of these currents is still an open question which may be related to the initial formation of the earth. This lack of understanding is, however, one of the principal reasons for obtaining measurements of the fields of other planets which formed under different conditions and have a different composition. Similar thoughts suggest that knowledge of the moon's magnetic field would be a major step toward determining how it formed as an earth satellite.

Direct measurements of the electric fields exterior to the earth's ionosphere and in interplanetary space are important to the eventual understanding of how cosmic rays and auroral particles are accelerated and from whence they come.

Program

Long-Range: Results from satellites, probes, and rockets to be flown in 1959 will be an important factor in determining the long-range program for studying the earth's fields. One can, however, anticipate that an important item will be establishing an earth satellite observatory which will include instruments for measuring particle flux and solar radiations as well as magnetic and electric field instruments such that direct correlations can be made between the various phenomena. Also rocket soundings into the ionosphere will continue to be used to study details of ionospheric currents more thoroughly.

The fields of the moon, Mars, and Venus will first be measured from probes making close approaches and eventually from packages landing and serving as observatories.

Probes will be shot at the sun to obtain solar field measurements as close to the sun as feasible.

In all measurements, except the rocket soundings, the experimental intent will be to obtain both the absolute magnitude and the direction of the magnetic field measured. It will be more difficult to separate the ambient electric field from that caused by an electric charge on the space craft.

Theoretical analyses and correlations between electric and magnetic field phenomena and other phenomena will be an integral part of the program.

Immediate: The short-range magnetic-field program includes:

1) a number of small sounding rockets with proton magnetometers for studying ionospheric currents,
2) two sounding rockets primarily for flight testing of newly-developed Rb-vapor magnetometers and for information on radiation belt currents,
3) a space probe to the moon with a Rb-vapor magnetometer for measuring electric currents and the form of the earth's field at great distances, interplanetary fields, and the moon's magnetic field.

The program also includes instrument development—the most significant being the development of Rb-vapor magnetometers by Varian Associates. It is also anticipated that simpler magnetometers which can detect only the existence of a perceptible field will be placed in several rockets and space vehicles as secondary experiments.

GRAVITATIONAL FIELDS

Objectives

To determine and understand the origins, natures, methods of propagation, spatial distributions, temporal variations, and effects of gravitational fields throughout the universe. To determine and understand the external form and internal constitution of the earth, planets, and stars. To determine and understand the relations between gravitational and electromagnetic fields. To evaluate effects of gravitational fields of different magnitudes, including weightlessness, on instrumented and manned space exploration.

Present Knowledge

At the present time, the best theory of the nature of gravitation is Einstein's general theory of relativity, which asserts that the force which we recognize as gravity results from a distortion of space-time. An essential part of this theory is the idea that mass as determined by gravitation will be found to be proportionate to mass as determined by inertia. Other theories have been proposed, especially by George D. Birkhoff.

There are three tests of the general theory of relativity. The first of these is the motion of the perihelion of Mercury, which amounts to 43 seconds of arc per year, agreeing quantitatively with Einstein's prediction. The second is the bending of light in the sun's gravitational field. It has been measured numerous times, and the results have generally been found to agree with Einstein's predictions; but, in view of the smallness of the effect (only 1.75 seconds of arc), there is still room for reasonable doubt about the confirmation.

The third effect is the shift of light toward red wavelengths in a strong gravitational field. This is doubtfully observed in the sun, where it is confused by mass motions of gases. A red shift is observed in some white-dwarf stars, but even there it cannot be said unequivocally that this is not an effect of mass motion.

It is further believed that the force of gravity is proportional to the density of matter in the universe. Since the universe is expanding, this means that changes of the order of one part in ten billion per year ought to exist in the force of gravitation.

With respect to the gravitational field of the earth, this problem would be solved fully if we knew all the harmonics into which the field can be subdivided. There is one zero harmonic which amounts to the product of the gravitational constant by the mass of the earth, and which has at the present time the value of 3.9868 with an uncertainty amounting to one part in 100,000. There are three possible first harmonics which arise from the displacement of the center of the earth from the center of our present coordinate system. This displacement is believed to amount, at the present time, to no more than 100 meters. There are five second-degree harmonics, of which one harmonic of zero order is intimately related to the flattening of the earth and has been measured with great precision from the motion of the node of 1958 Beta 2 (Vanguard I). Coefficients so obtained correspond to a flattening of 1/298.3. Two of the remaining four are necessarily zero; and the other two have been doubtfully evaluated with an amplitude of about 4 mgals.

Of the seven third-degree harmonics, the zonal harmonic of zero order has very recently been measured, and appears to have a value of 5.6 mgals at the surface of the earth. The other six have been estimated crudely from gravity data. An estimate of 5 mgals has been made for the fourth zonal harmonic. It is known that the size of these harmonics is proportional to the stresses at the interior of the earth roughly at the rate of 1 mgal to 10 kg per square centimeter.

In the case of Mars, the zero harmonic (the mass) is 0.1076 that of the earth. Nothing is known with respect to first harmonics. The second harmonic is determined from the flattening which is found to be 1/192 from the satellite motions, and 1/77 from visual observation. Masses and flattenings for Jupiter, Saturn, Uranus and Neptune are available in the same way, but no higher harmonics are known for any of these planets.

The sun's mass is accurately determined by the motions of the planets. Nothing is known about any other higher harmonics.

Measurements of various double stars yield masses which are found to fall into the range of from one tenth to fifty times the solar mass. The flattening, which ranges from zero up to the limit of instability, is known for a few eclipsing binary stars.

The gravitational field of the galaxy as a whole has been deduced from the rate of galactic rotation. It is found to be equivalent to several hundred billion solar masses. The gradient of the galactic field which measures the flattening of the galaxy has also been determined from the gradient of galactic rotation near the sun.

The nature of the intergalactic gravitational field is at present completely unknown; it is not even certain that a replusion does not exist at sufficiently great distances.

Existing Problems

With respect to the fundamental nature of gravitation, the principal problem is to find some peculiarity of the law which would serve to control the formulation of satisfactory theories. Such a peculiarity might consist, for example, in a small deviation from strict equality of gravitational and inertial mass, or in a velocity of propagation, or in something in the nature of a shielding such as is observed with magnetism. Any one of these effects would give us a clue to the best direction to look for a more fundamental understanding.

Within the framework of the Newtonian Law, the classic problem is that of the long-time stability of a system such as the solar system. The problem may be formulated in this way: As we think of time of the order of 10^9 years, do the orbital planes of the planets tend to approach one another? Do the eccentricities tend to decrease? If so, then, of course, we understand the architecture of our present solar system; if not, then we have an important clue which can someday be used to discover how it was initially formed. The problem of the stability of the solar system has engaged some of the best mathematicians since the days of Lagrange and Laplace, who found that stability could be guaranteed for times of the order of 10^8 years.

With respect to the gravitational fields of the planets, one problem is that of the presence of iron cores in the planets. At the present time, the large density (approximately 5) found for Mercury indicates a large iron core. It is hard to put this in an orderly relation with the density of Mars (about 3.9) and the moon (about 3.3). In particular, if correct, this density means that the iron core found in some planets cannot represent a pressure modification of ordinary silicate rock. A second test of the pressure modification idea is furnished by the density of Venus, which is just at the critical level. In both cases, it appears possible that further observational data would reverse the conclusions which have been drawn.

The gravitational field of the earth is closely related to a series of interesting problems. In the first place, the zero harmonic is essential for the study of space trajectories. Next, the internal constitution of the earth manifests itself through variations in the higher harmonics. In particular, a state of hydrostatic equilibrium in the interior of the earth demands the value of 1/297.3 for the flattening of the earth. The discrepancy between this and the observed value of 1/298.3 is a major problem in determining the structure of the earth. If the discrepancy is due to mechanical strength in the interior, then it may be related to the observed large values of some of the higher harmonics.

At the present time, two independent and contradictory theories have been offered for the inner structure of the earth and for the formation of mountain ranges. The first of these, which we might call the plastic theory, envisions a viscous interior in which slow convection currents of a few centimeters per year exist, and mountain ranges, such as those around the Pacific, are formed as the result of drag on the underside of the continents. The other theory, which we may call a crystalline theory, considers that the earth is slowly contracting as it cools, and attributes the formation of mountains to crumpling. At the surface of the earth, the problem of verifying or disproving the "basic hypothesis of geodesy," namely that the product of gravity anomalies in milligals times areas in square kilometers does not exceed 30 for any extensive patches, is one of the most important that can be set up.

Program

Long-Range: For the study of the fundamental nature of the gravitational field, two avenues are opened by our ability to launch satellites and space probes. The first of these is the ability to try experiments on a scale of hundreds and thousands of kilometers by probing the fields of planetary masses with bodies capable of being accurately observed. This is significant because gravitational fields, except that of the earth, are almost unmeasurable in laboratory-scale experiments. The second avenue results from the fact that the periods of artificial satellites are so much shorter than those of the moon and other natural satellites that we can observe in a few years a number of revolutions corresponding to thousands of years for natural satellites. By the first avenue, we hope to find the links which must exist between the theory of the electromagnetic field and the gravitational field. It is planned, in particular, to test the equality of gravitational and inertial masses by experiments in space on a colossal scale, which are a repetition of the experiment of Galileo in the Leaning Tower of Pisa. An attempt will be made to devise experiments which will reveal the velocity of propagation of gravitation, if any.

It is planned to determine the masses of the inner planets by direct observation of probes passing near them or possibly around them. These probes will, at the same time, help to determine the value of the astronomical unit, the fundamental meter-stick of the solar system.

It is planned to test the hypothesis that gravitational attraction depends on the average density of matter in the universe and that, therefore, it is slowly weakening as the universe expands. For this purpose, an atomic clock on the ground can be compared with a gravitational clock of some kind. A proposal for a gravitational clock consisting, in effect, of a high satellite with a very well measured orbit is being studied.

It is planned to employ moon probes to obtain improved values for the over-all mass of the moon and for the moments of inertia about its three principal axes. A determination of the strength of the materials in the moon's interior will be attempted from this information.

It is planned to measure the mass of Venus and of Mercury in order to test Bullen's idea about the nature of the cores of the planets.

Using the second avenue, it is planned to observe the motions of close satellites of the earth over a long period and to make a precise comparison with theory, searching for systematic trends in the inclination and the eccentricity, which might shed light on the history of the solar system.

Immediate: There are several phases of the short-term gravitational field program:

1) Studies are now being made on existing satellites with the object of determining the low harmonics of the earth's field from tracking data.

2) Preparations are made for equipment for a special geodetic satellite which will be capable of refining the observations on the harmonics, and of determining intercontinental distances with high precision. It will clearly be possible to carry the study of the form of the geoid much further than has been possible to date.

The information developed in 1) and 2) above will be applied to the question of the basic hypothesis of geodesy. This hypothesis, as formulated by some theorists, is in essence that the low harmonics of the earth's gravitational field have amplitudes of a meter or so. The hypothesis is not universally accepted; other theorists consider that the amplitudes are of the order of scores of meters. A decision between these two hypotheses is important because Heiskanen, in particular, proposes extensive work based on Stokes' theorem—work which is warranted only if the basic hypothesis is satisfied. This information will also be used in an attempt to evaluate hypotheses of convection in the mantle. These hypotheses seem to go with the ideas of the first-mentioned school of theorists, and it may be possible to decide between these hypotheses and the alternative contraction hypotheses on the basis of our information.

3) It is planned to put in orbit a satellite carrying a very precise clock in order to test a theory of Einstein which predicts a change in the clock's speed with a change in the strength of the earth's gravitational potential. If successful, this would constitute a new test of the general theory of relativity, and so improve our understanding of the nature of the earth's gravitational field.

ASTRONOMY

Objectives

To determine the spatial distributions of matter and energy over the entire universe, and to understand their cosmological origins, evolutions, and destinies. To observe from above the earth's atmosphere the spectral distributions of energy radiated from objects in the solar system, in this and other galaxies, and in the intervening space, with emphasis on observations that are prevented or compromised by the absorption, background emission and refraction of the earth's atmosphere. To determine and understand the structure and the surface features of the planets. To determine the effects of meteors, radiations, and other astronomical influences on instrumented and manned space exploration.

Present Knowledge

The field of astronomy may be divided broadly into five categories: 1) the solar system, 2) the stars, 3) interstellar matter, 4) the characteristics of stellar aggregates, such as clusters and galaxies, 5) the history and evolution of the universe and its various components.

The Solar System: The dark bodies orbiting around the sun may be classified as planets, satellites, minor planets or asteroids, and comets. The nine planets, including the earth, differ primarily in size and in temperature. The smallest ones, such as Mercury, have too little gravitation to retain any atmosphere, while the larger planets, such as Jupiter and Saturn, have extremely dense atmospheres. Those planets closest to the sun have, of course, quite high temperatures, whereas the most distant planets are very cold. With respect to the amount of atmosphere and to temperature, Venus, Earth and Mars are intermediate, and these three planets are considered to be the only ones that might naturally support any form of life. Earth may be the only planet on which conditions are so ideal as to favor the evolution of intelligent life. Asteroids and most of the satellites of planets are small bodies, ranging from dust particles to boulders a few hundred miles across. The comets are loose agglomerations of rock and gas that spend most of their time in the far reaches of the solar system and approach the sun at infrequent intervals. The asteroids and comets are probably the debris left over from the formation of the planets some five billion years ago.

The Stars

It is now known that the stars are gaseous bodies composed of about 75 per cent hydrogen by weight, 25 per cent helium, and traces of the remaining elements. Most stars derive their energy from the conversion of hydrogen into helium deep in the stellar interior where temperatures and pressures are extremely high. This energy is transferred to the outer layers of the star by radiation and convection; there the emerging light is affected by the constituents of the stellar atmosphere, producing the observable stellar spectrum. This spectrum, together with independent information concerning the star's luminosity and mass, is the basic datum from which all knowledge of the star's structure must be deduced. A significant stellar characteristic is the fairly well defined correlation between surface temperature, as evidenced by a star's spectrum or color, and intrinsic luminosity. The form of this temperature-luminosity diagram has yielded much information concerning the nature of stellar atmospheres. From the study of spectra it is found that the chemical constituents are approximately constant from star to star; however, other parameters vary widely. Surface temperatures range from 1500°K to more than 50,000°K. Stellar diameters vary from less than 0.01 to more than 500 times that of the sun; masses range from 0.1 to 50 solar masses; and intrinsic luminosities vary by a factor of a hundred million. The outer layers of a star are far from being in equilibrium. This is obvious in the case of the sun, where sunspots and their accompanying magnetic fields, intense hot spots called flares, and eruptions from the surface, called prominences, indicate large-scale thermal and magnetic instabilities. The cause of these instabilities and, more basically, the reason for their twenty-two year periodic variation are not yet understood. Other stars show even more striking examples of instability; stars which pulsate with periods of days or even hours; others which exhibit large, highly-variable magnetic fields; and, finally, the novae and super-novae, which are subject to catastrophic explosions.

Interstellar Matter: The material between the stars has a composition similar to the stellar composition, in the form of individual atoms and radicals and small dust particles, or grains. The space density of this material is a few atoms per cubic centimeter and except near hot stars it has a kinetic temperature of about 100°K. In spite of the low density, the total mass of interstellar matter in the Milky Way is approximately equal to the total stellar mass. This material has a profound effect on stellar observations because its scattering reduces the intensity of starlight passing through it, and because it superposes interstellar absorption lines on stellar spectra. In addition, the material has a marked tendency to occur in clouds near hot, luminous stars, which themselves are often found in clusters. In these cases, the interstellar clouds are illuminated and become visible as reflection nebulae. If the nearby stars are hot enough to excite the gas, an emission nebula results, emitting energy primarily in the lines of hydrogen and of some of the trace constituents. In these nebular complexes, the density may be as high as a thousand atoms per cubic centimeter, and the kinetic temperature is usually about 10,000°K. Knowledge of the interstellar medium has increased rapidly since the development of radio astronomy. Radio-frequency emission has been detected from the neutral hydrogen in the insterstellar gas, and has proved to be a valuable tool in probing distant parts of the galaxy. Thermal RF radiation has been observed from emission nebulae, and RF synchrotron radiation has been measured from nebular remnants of supernovae explosions. The polarization of starlight has indicated the existence of magnetic fields in the interstellar medium which may play an important role in the production of cosmic rays.

Star Clusters and Galaxies: Stars tend to occur in multiple systems of two to six stars and in clusters of tens to many thousands. The study of such clusters has been valuable because the members presumably have common origins and histories and thus present homogeneous populations suitable for statistical analysis. On a larger scale, stars are clustered into galaxies containing billions of members. In appearance, these galaxies range from the smaller systems, which are spherical, ellipsoidal or irregular, and composed of fainter stars and little interstellar gas, up to the massive, highly-flattened spiral galaxies which show a great deal of complex structure. Our galaxy, the Milky Way, is an example of the latter type. It consists of a central spheroidal nucleus of stars, surrounded by a round stellar sheet, quite thin compared to its diameter. Within this sheet, the galactic plane, the brightest stars, and most of the interstellar material are concentrated into spiral arms. The sun is within such an arm. However, in addition to these stars, there is a second group of stars having different temperature-luminosity and velocity characteristics which are spherically distributed about the galactic nucleus. These so-called Population II stars are quite similar to the stellar populations of the smaller elliptical galaxies. It appears that spiral structure with its associated interstellar material and bright stars may be a superficial and rather recent phenomenon superposed on an older and more fundamental stellar population of which all galaxies are composed.

Stellar Evolution: With the possibility in mind that there are stars with different ages and histories, the differences between the temperature-luminosity diagrams of different groups of stars, and especially of different star clusters, may be interpreted as being due to age and evolutionary differences as well as to small variations in chemical composition. The presently available evidence points to the following picture of galactic and stellar evolution. The original material of the universe coalesced into discrete globs or proto-galaxies and, in turn, stars and star clusters condensed out of these globs. After the initial period of star formation, those galaxies which had no excess material, or which lost their excess material through intergalaxy collisions, evolved into the elliptical galaxies. In the remainder, the excess material collapsed into flat sheets in which star formation has continued at a slower pace up to the present time. Star formation can occur when a local inhomogeneity in a dense interstellar nebula becomes large enough and dense enough to contract due to self-gravitation. The interior pressure and temperature eventually becomes sufficiently high to support thermonuclear reactions and the gas becomes self-luminous, after which radiation pressure prevents the accumulation of appreciable interstellar material. The size and luminosity of the new star are determined by the amount of material accumulated before energy production begins. Multiple star formation from a single cloud is common. In some cases, the nebular material remaining around a new star condenses into planets. Single stars not surrounded by planetary systems are probably rare. Stars consume themselves by an ever accelerating process of converting hydrogen into helium until the hydrogen is exhausted. At this point the internal support of the star due to radiation pressure is removed and the main body collapses, while a tenuous outer atmosphere expands to tremendous proportions to preserve dynamic balance. The main body contracts until the central temperature and pressure become high enough to support the conversion of helium into heavier elements with an accompanying release of energy. During this evolution, the star may pass through phases of dynamic instability where pulsation occurs. Eventually the star has no more energy resources left to support its mass and an implosion occurs, producing a nova. The post-nova stellar core may then settle down into an extremely dense sub-luminous white dwarf star until it goes out completely.

Current Problems

The Sun: The sun is the only star on which we can study small regions of the surface and atmosphere. For this reason, and because solar radiation profoundly affects the physical conditions in the earth's atmosphere, the sun is an object of prime importance for detailed study. To gain a better physical picture of the conditions in the undisturbed sun, more accurate measurements are needed of the solar constant and of the energy distribution in the entire solar spectrum, not just in the region observable from the surface of the earth. Also, particularly because of its effect on terrestrial conditions, a long-term study of the disturbed sun in all wavelengths, especially in very short optical wavelengths and in very long radio wavelengths, is of great importance.

The causes and nature of the phenomena occurring in the solar atmosphere and their effects on physical conditions in interplanetary space are of vital importance in the climatology of the planets. Some basic problems involve the propagation of disturbances through the solar atmosphere, the role of convective transport of energy in this atmosphere and the origin, support and heating of the chromosphere and corona.

The Solar System: There are still many pertinent questions concerning the physical characteristics of the planets and the other constituents of interplanetary space. More information is required on the percentage concentration of oxygen, water vapor, and other important gases in planetary atmospheres; the physical properties of the planetary surfaces; and the structure of planetary interiors. Since the maximum intensity of the thermal radiation from the planets occurs in the infrared, it is obvious that studies in these wavelengths are important for deriving the physical characteristics of the planetary surfaces. The important band spectra from planetary atmospheres tend to occur in the infrared also. In addition, even those absorptions in the region of the spectrum accessible from the earth's surface are partially obscured by similar molecules in the terrestrial atmosphere. Finally, the information needed on the structure of planetary interiors can best be derived by studying the effects of these planets on nearby satellites. Artificial satellites which should eventually prove particularly useful for this purpose.

An interesting question is that of the existence of a permanent interplanetary gas. If this exists, is it primarily the interstellar gas through which our solar system is moving or is it matter ejected from the sun or part of the solar atmosphere? The existence of such a nebula would provide a laboratory for studying general nebular properties, and for studying at first hand a gaseous nebula surrounding a star. In addition, the existence of a solar nebula may also limit our study of nebulae elsewhere in the stellar system.

Stellar Physics: More accurate information is needed on fundamental quantities such as stellar masses, radii, and luminosities. For stars much hotter or much cooler than the sun, the luminosities must be extrapolated from the very small spectral region accessible from the surface of the earth to the regions which contain the predominant energy output of these objects. In the field of stellar atmospheres, more information is needed on the relative importance of radiative transfer as compared to the transfer of energy by mass motion, and on the origin and effects of stellar rotation and magnetic fields. Important information on these problems is provided by the relative energy output in widely separated regions of the stellar spectrum. Our current knowledge of stellar composition, structure, and evolutionary processes depends upon spectroscopic observations in a relatively limited region of the spectrum. It is obviously desirable to extend our observations over the entire electromagnetic spectrum. These observations are especially important for the hot stars which are in rapid evolution. Both their maximum energy output and the spectral lines which are critical for studying the abundances of important elements in these stars are in the currently-inaccessible ultraviolet.

Interstellar Medium: In the study of the interstellar medium, the question of the energy balance has been reopened by recent rocket investigations, which show more ultraviolet energy being released by radiative processes than had been previously supposed. Additional information is needed regarding the nature of interstellar grains, their absorption and scattering of starlight in a wide range of wavelengths, particularly in the infrared, and their interaction with the interstellar magnetic fields. In studies of the interstellar gas, astronomers have been handicapped by the lack of observation in the ultraviolet portions of the spectra where the principal spectral lines of the interstellar gases lie. These data promise to supply vital information on the chemical composition, the distribution relative to galactic features, and the mass motions and excitation processes which play an important role in the evolution of our galaxy.

Star Clusters and Galaxies: Star clusters, galaxies, and clusters of galaxies constitute the large-scale building blocks of the universe. The role and mechanism of these associations remain questions basic to the solution of the origin and evolution of our universe. The individual properties of the star cluster and the galaxy such as mass distribution, mass-luminosity ratio, angular momentum, gas and dust concentration, and energy balance are typical of the parameters needing study. At present, we have little knowledge of the nature and density of the material in intergalactic space, as observations in the visible spectrum have proved inconclusive. It is desirable to carry the observations into ultraviolet portions of the spectrum, where higher absorption can be expected to greatly increase the sensitivity of these measurements. Current theories in cosmology depend strongly upon the correct interpretation of the red shift. It is desirable to determine if this red shift is a constant over the whole electromagnetic spectrum, and also if it is a linear function of distance. Since the red shift for very distant galaxies is sufficient to shift the normally-observed spectral features into the unobservable infrared, detailed studies of the ultraviolet region of nearby galaxies and the infrared of very distant galaxies are essential to the detection of possible evolutionary effects during the billion or more years it has taken the light from the further galaxies to reach the earth.

Program

Long-Range: The great distances of astronomical objects outside of the solar system mean that they are both small in angular dimensions and apparently faint. Hence, for wavelengths shorter than one meter, the objects require long periods of observation with large radiation collectors having good angular resolution and accurate pointing. Radio-astronomical observations in wavelengths beyond the ionospheric limit require extensive antenna arrays to achieve good resolution. Thus, until an observatory on the moon is possible, experiments in this wavelength region will be limited to the sun and to low-resolution reconnaissance surveys. In the short-wave radio regions, astronomical sources are weak. It is possible that the development of large inflatable balloons will provide the necessary collecting area for observations in this region. To provide the necessary pointing accuracy and observing time for experiments in the infrared, optical, ultraviolet and X-ray regions of the spectrum, development has begun on an orbiting stabilized platform. It will be possible to point this platform toward any point in the sky and to hold it in that direction for periods from a few minutes to an hour, with sufficient accuracy to allow continued observations of individual celestial objects. The satellite will contain instruments of moderate size which can be completely controlled from the ground, thus providing a remotely operated observatory which can perform most of the functions of a ground-based observatory. Although the sky above the earth's atmosphere will be dark, the presence of sunlight scattered within the vehicle itself may be disturbing. The eventual establishment of an astronomical observatory on the moon will eliminate this difficulty, since the moon itself can provide an effective shield against the sunlight. Moreover, the establishment of an observatory on the moon will undoubtedly make feasible larger equipment than can be placed on a satellite-borne platform. The execution of these programs will require the development of not only a stabilized platform and its pointing controls, but also specialized optical equipment and radiation receivers for particular wavelength regions, especially adapted for a particular program, and improvements in telemetering techniques so that transmission to ground stations of spectroscopic data of adequate scope and resolution will be possible. For planetary studies, the planetary satellites and planetary probes planned for the atmospheres program will also be useful for providing information on the surface conditions and interior structures of the planets.

Immediate: The immediate program includes the exploration of the moon; reconnaissance measurements in the ultraviolet region of the spectrum to determine the brightness and positions of interesting regions of the sky; preliminary testing of equipment and techniques to be used later on the stable platform; and studies of the sun in the ultraviolet and X-ray region with pointing controls somewhat less stringent than those to be used for the stable platform. The reconnaissance will continue and extend to the southern sky the survey of the newly-discovered extended emissions in the far ultraviolet in order to determine the nature and sources of these emissions. Stellar photometry measurements will be made in the near and far ultraviolet spectral regions to extend the magnitude systems to these wavelengths. Emphasis is being given to observations in the previously unexplored far ultraviolet and high-energy gamma-ray spectral regions. Studies of the solar ultraviolet and X-ray spectra will include their long-term variations, line profiles, distribution across the disk, and the spectra of the coronal X-ray emissions. In addition, long-wavelength radio receivers will be placed in satellites to study the brightness distribution of the sky in these wavelengths, the spectrum of the sun and its variation, and the absorption of the earth's ionosphere at various heights.

BIOSCIENCES

Objectives

To determine the effects on living terrestrial organisms of conditions in the earth's upper atmosphere, in space, and in other planetary atmospheres. To determine the effects of flight through space on living organisms. To investigate the existence of life throughout the solar system, and to study in detail whatever extraterrestrial life forms exist.

Problems

The opportunity now exists to conduct fundamental life-sciences research in satellites and space probes. In such research, conditions in space and conditions during flight through space enter as new experimental variables. Principal interest lies in the components of the space and space-flight environment that are irrevocably different from the terrestrial, such as radiation and the altered gravity state. The former is qualitatively and quantitatively different from terrestrial radiation and looms as a far more important problem than originally thought. The latter offers the possibility of elucidating biological processes known to be effected by weight or apposition of one part against another, such as cell division and organogenesis. These and other areas of study could be carried out in satellites or space probes from which specimens under study could be recovered.

More specifically, fundamental experiments on living organisms in the satellite environment or in other types of space vehicles and stations should be conducted to cover the following general topics: 1) Animal navigation and orientation, 2) Mitosis and embryology, 3) Plant morphogenesis, 4) Geotropic response of plants to altered gravitational fields, 5) Biological rhythms and cycles, 6) Altered gravitational effects on blood circulation in animals, 7) Vestibular physiology, 8) Neuropharmacology—effects of tranquilizers, motion sickness drugs, etc., 9) Effects of cosmic radiation, 10) Tolerance limits for combined stresses, 11) Physiological and psychological deterioration, 12) Effects of acceleration and deceleration.

There has been considerable speculation about the possibility of the existence of extraterrestrial life forms. The debate ranges from whether or not life exists on Mars and Venus, and if so in what form, to whether or not spores and similar life forms can survive the rigors of interplanetary space. Investigations of these topics must, for the most part, await observations which can be made on direct samples and studied in space laboratories of the future. It is conceivable, however, that with proper and specialized instrumentation and remote sampling techniques it may be possible to obtain evidence of primitive or complex organic material without the presence of a scientist and a laboratory in the sample area. Elaborate instrumentation and the recovery of records, or advanced telemetering techniques could provide information about the nature of biochemicals on the planets.

Program

The program in biosciences is still being formulated. Immediate steps include a series of discussions with leading bioscientists to firm up important initial lines of attack.

NASA and Industrial Property*

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Summary—**The purposes and powers of the National Aeronautics and Space Act are described briefly, particularly as they affect the electronics industry, as an introduction to a more extended discussion of the provisions of the Act relating to patent rights and rights to trade secrets and proprietary information generally. The stringent nature of the latter provisions is stressed, and adverse effects of strict construction and enforcement on the electronics industry analyzed. Mitigating regulations under which the Act will in fact be administered are discussed.**

INTRODUCTION

THIS PAPER concerns itself generally with the National Aeronautics and Space Action of 1958 (Public Law 85-568), hereinafter referred to as the Space Act. More particularly, we shall be concerned with the effect of the Space Act on industrial property, such as patent rights, confidential information, and trade secrets. We shall endeavor to derive the legal, business, and economic consequences of the Space Act as these might flow from the terms of the Act itself, and from the interpretations of the Act being accorded to it by its current administrators.

The Space Act establishes a civilian agency, called the National Aeronautics and Space Administration (NASA). The Act provides for an Administration, and for a National Aeronautics and Space Council which exercises high level or policy control of NASA. The latter is presided over by the President of the United States, and includes among its members the Secretary of State, the Secretary of Defense, the Administrator, the Chairman of the Atomic Energy Commission, and others. Its basic function is to set policy for NASA. In terms of corporate organization, NASA has, as Chairman of its Board, the President of the United States, and on its Board of Directors some of the highest officers of the United States Government, although *pro forma* the Council is not part of NASA. The Council is also a coordinating agency between NASA, DOD, and scientific agencies of the Government.

A basic function of NASA is to conduct "aeronautical and space activities" sponsored by the United States, except those activities which are "peculiar to or primarily associated with the development of weapon systems, military operations or the defense of the United States." The exclusion extends to research and development "necessary to make effective provision for the defense of the United States." These latter functions remain the responsibility of the Defense Department.

The term "aeronautical and space activities," which constitute functions of NASA, are very broadly defined in the Act, and NASA is not restricted by the terms of the Act to activities having a direct or necessary relation to outer space. NASA personnel insist that the broad language of the Act was intended merely to minimize restriction on activities of NASA, without any intention of involving NASA in activities not concerned with outer space. Despite these protestations, the Act itself provides that, apart from defense activities, NASA has cognizance of aeronautical as well as space activities and that its objectives, in these fields, are 1) to expand human knowledge of phenomena in the atmosphere and space, 2) to improve the usefulness, performance, speed, safety, and efficiency of aeronautical and space vehicles, 3) to develop and operate vehicles capable of carrying instruments, supplies, and living organisms through space, 4) to establish long-range studies of the benefits, opportunities, and problems involved in aeronautical and space activities, 5) to preserve the United States as a leader in aeronautical and space science and technology, 6) to convey information concerning its accomplishments to defense agencies where these accomplishments have military value or significance, 7) to receive from the military agencies all information which may have value or significance to NASA, 8) to cooperate with other nations in peaceful application of aeronautical and space activities, and 9) to promote the most effective utilization of scientific and engineering resources of the United States, with close cooperation among all interested agencies of the United States, in order to avoid unnecessary duplication of effort, facilities, and equipment.

The Act defines "aeronautical and space activities" as meaning 1) research into and solution of problems of flight within and outside the earth's atmosphere, 2) development, construction, testing, and operation for research purposes of aeronautical and space vehicles, and 3) any other activities which may be required for the exploration of space. The Act defines "aeronautical and space vehicles" as meaning aircraft, missiles, satellites, and other space vehicles, manned and unmanned, together with related equipment, devices, components, and parts.

It immediately becomes apparent that the purpose of the Act, as stated in the Act itself, is not restricted to space activities. NASA is empowered, and in fact directed, to concern itself with aircraft, and with related equipments, devices, components, and parts, and to improve the usefulness, performance, speed, safety, and efficiency of aeronautical vehicles. It follows that NASA may, within the terms of its authority, concern itself with such things as telemetering, communication systems, guidance systems, radar systems, navigational systems, and, in terms of devices and components, may concern itself with the entire electronics industry.

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A further function of NASA which has not received attention heretofore relates to point 9); *i.e.*, that NASA is to so conduct the aeronautical and space activities of the United States as to contribute to the most effective utilization of scientific and engineering resources of the United States in order to avoid unnecessary duplications of effort, facilities, and equipment. Since the purpose of the Act in relation to aeronautics, space activities, and any adjunct thereto, is extremely broad, it is not far afield to state that the Act provides authority to regiment the scientific and engineering resources of the United States. Avoidance of the duplication of effort in telemetering, for example, or in radar, or in radio communications or the avoidance of unnecessary duplication of effort, facilities, and equipments in these fields, imply powers having far reaching implications. This is not to say that the power will be utilized to the extent for which the Act provides authority, but merely that the authority exists and that the Council is of sufficiently high level that if NASA decided to exercise the authority granted to it, there would be no one to intervene unless Congress should decide to modify the Act or to repeal it.

The Act provides that it shall be the duty of the President, as a member of the National Aeronautics and Space Council, to survey all significant aeronautic and space activities of all agencies of the United States engaged in such activities. Obviously, this includes the Defense Department. The Act further orders the President, as follows: He shall develop a comprehensive program of aeronautical and space activities to be conducted by agencies of the United States, and designate and fix responsibility for the direction of these activities. He shall provide for effective cooperation between NASA and the Defense Department in respect to such activities, and specify which activities may be carried on by both agencies, notwithstanding assignment of primary responsibility to one of these. The latter provision obviously provides for concurrent jurisdiction of defense matters in NASA and the Defense Department should the President decide that this is necessary or advisable. Moreover, the President is empowered to resolve differences among the departments as to whether a particular project is or is not an aeronautical and space activity. Since aeronautical and space activity is so broadly defined as to include almost any conceivable activity in the field of science, enormous vistas are opened up.

PATENT PROVISIONS

We have reviewed the purposes, functions, and powers of NASA, as provided by the Space Act, as a prelude to discussion of the patent provisions of the Act. Were these powers and functions restricted to the field of space activities, *i.e.*, to satellites, space platforms, interplanetary travel, and the like, provisions of the Act relating to patents and to industrial property generally would not be of primary interest to the greater part of the electronics industry. However, since the Act is so broadly worded that there is no field of science in which its voice may not be heard in due course, and since there are few scientific functions of the Defense Department which NASA may not absorb or parallel eventually, the patent provisions and the industrial property provisions of the Act acquire great significance.

Section 305 of the Act provides that "whenever any invention is made in the performance of any work under any contract of the Administration and the Administrator makes certain determinations, the invention and patents predicated thereon shall be the exclusive property of the Government." This in itself appears to be similar to provisions found in the Atomic Energy Act, but involves a policy which is not followed by the Department of Defense. It becomes pertinent, then, to consider the meanings of the terms "invention," "made," "performance," "work," and "contract." The act is explicit in defining some of these. For example, the term contract is defined in Section 305(j)(2) of the Act, as follows: "The term contract means any actual or proposed contract, agreement, understanding, or other arrangement, and includes any assignment, substitution of parties or subcontract executed or entered thereunder."

One who enters into a contract may, as a general proposition of law, agree to forego rights to patents which may arise in its performance. It is his privilege to refuse to enter into the contract if he does not wish to forego his rights to patents. A departure from previously known practices arises in the words "proposed contract." The term can well be construed to mean that one who submits a proposal to NASA, thereby and without a contract or any understanding, or any consideration of any kind, transfers rights to patents to NASA. A further area worthy of thought relates to the term "subcontract," which is not defined. Is a supplier to a NASA contractor *per se* a subcontractor, and thereby one who performs work under a contract with NASA?

We may envisage, for the sake of example, a course of action in which a company interested in electronics makes a proposal to NASA, the proposal being sufficiently vague that NASA may refuse to consider that it includes a conception of an invention. The company is now a "contractor" by virtue of having made the proposal. It is then given some encouragement to proceed by NASA, at some expense to itself, to develop its proposal to completion in respect to means and methods. It soon has a complete conception of an invention, which it has completed while a contractor. NASA is pleased with the proposal but decides to award a contract to perform the work thereunder to a competitor, on the ground that the latter is better qualified for the purpose. Under the terms of the Act, NASA can take full right, title, and interest to the invention so made, and refuse to compensate the contractor for his efforts. It is true that under the terms of the Act the Administrator is not required to take full right, title, and interest, as we shall see, but he is required to take at least a royalty-free licence.

The term "made" when used in relation to any invention, means 1) conception, or 2) first actual reduction to practice. One who merely conceives of an invention and who does no more can thereby have forfeited all rights to the invention to NASA if he is an NASA contractor. One who conceives of an invention prior to making a proposal thereon to NASA has made an invention prior to such proposal, and at that point the invention is his own. On submitting the proposal he has entered into a contract, but since the invention was made prior to entering into the contract, the inventor may retain his rights. If the inventor actually reduces to practice following the proposal, without an NASA contract, and certainly if he receives a contract and thereafter reduces to practice, he forfeits his rights to the invention, because he has reduced to practice while a contractor.

The Act does not define the word invention, but in specifying what reports shall be provided under NASA contracts, provides for "a written report concerning any invention, discovery, improvement, or innovation." This would appear to go further than does the Defense Department. An innovation may be a very slight thing, whereas an invention goes beyond what can be expected of a man skilled in the art in which he is working. Furthermore, the report is required to contain "full and complete technical information," a requirement which should be considered in relation to the further requirement that NASA disseminate the information which it collects. The Defense Department has not heretofore generally required full and complete technical information concerning inventions, but has usually been satisfied with sufficient information to enable a patent application to be prepared. There is a great distinction between information which is sufficiently detailed to form a basis of a patent application, and full and complete technical information. The latter may well be interpreted to mean complete design information, or manufacturing information. If NASA makes a practice of acquiring full technical information and disseminating the same, a great mass of heretofore proprietary information may eventually become public.

Section 305 provides that inventions made in the performance of work under a contract with NASA shall be the property of the United States, and that if such invention is patentable, a patent shall be issued to the United States upon application made by the Administrator. These words, if taken at face value, might indicate that the Administrator is to apply for patents in his own name. This may be merely lax language in the statute, since heretofore under United States law the primary applicant for Letters Patent has been the inventor or inventors, unless he or they were dead or mentally incompetent. In the case of the death of an inventor, his executor or the administrator of his estate may file in his name, and in the case of an insane inventor, his guardian may file. There is, however, one additional situation in which a person other than an inventor may become an applicant for patent. Rule 47 of the Rules of Practice of the U. S. Patent Office provides that "whenever an inventor refuses to execute an application for patent, or cannot be found or reached after diligent effort, the person to whom the inventor has assigned or agreed in writing to assign the invention, or who otherwise shows sufficient proprietary interest in the matter, justifying such action, may make application for patent on behalf of and as agent for the inventor." We may, therefore, find, in cases where conflict arises as to whether or not an application is to be filed, or to whom it belongs, that the Administrator will, pursuant to Rule 47 of the Rules of Practice in the United States Patent Office and pursuant to the authority provided in the Act, file on behalf of the inventor, and as his agent.

Section 305 provides for certain fact determinations to be made by the Administrator, upon the basis of which he may apply for a patent, covering work done under an NASA "contract." The contractor may dispute such fact findings. In summary, if the Administrator finds that 1) the person who made the invention was employed or assigned to perform research, development, or exploration work and the invention is related to the work he was employed or assigned to perform, or that it was within the scope of his employment duties, then whether or not the invention was made during working hours or with a contribution of the Government, or the use of Government facilities, equipment, materials, allocated funds, information proprietary to the Government, or services of Government employees during working hours, the patent belongs to the Government; 2) where the person was not employed to invent or to perform research and development or exploration work, but the invention is nevertheless related to the contract, or to work or duties he was employed or assigned to perform and was made during working hours, or with a contribution from the Government of the sort referred to in situation 1) above, NASA is authorized to seize the patent rights involved, and, in the terms of the Act, the Administrator "may apply for a patent."

In essence, under situation 1), inventions of research and development personnel working under a NASA contract almost inevitably belong to the United States so long as the invention is related to the work of the contract, even if it is made on the inventor's own time, with his own money, and with no contribution from the Government. Situation 2) goes beyond situation 1) in that any invention made by personnel not hired to do research and development, and working under an NASA contract, belongs to the United States if the work is related to the contract, but then only if the work was done during working hours, or with a Government contribution of one of the types specified in situation 1).

Considerable confusion may arise in respect to NASA contractors who have personnel working in part on NASA contracts and in part on commercial or Defense Department contracts. This may readily occur in respect to supervisory personnel and consultants. We need only envisage, for example, a telemetry group with one NASA contract, one Defense Department Contract, and an allocation of company money, working on three subjects, all relating to telemetering. The group supervisor makes an invention during working hours, of primary application to the Defense Department contract. The invention is certainly related to the work on the NASA contract, since both relate to telemetering. Is the company subject to the Space Act, in respect to that invention, or can it retain commercial rights as normally provided in Defense Department contracts?

Under Section 305(c) of the Act, the Commissioner of Patents is required to refuse to issue patents which he believes have "significant utility" in the conduct of aeronautical and space activities, unless an applicant on request files a statement under Oath, setting forth the full facts concerning the circumstances under which such invention was made and the relationship of the invention to NASA contracts.

After the statement under Oath is filed, the Administrator is given 90 days to request that the patent be issued to the United States. If he does not do so, the patent issues to the applicant. If he does so, however, the applicant may contest his findings by filing a request for a hearing before a Board of Patent Interferences, which acts as an appellate tribunal in this respect. From the determination of the latter each party to the controversy may appeal to the Court of Customs and Appeals.

Even if the Administrator takes no action within the 90 days allowed, and the patent issues to the applicant, the matter may be reopened by the Administrator within five years, if he has "reason to believe the statement filed by the applicant contained any false representation of any material fact." Note that fraudulent representations are not required. Innocent but false representations are adequate.

The Administrator is authorized to waive his right to acquire full rights in inventions, for the United States, if he determines that "the interest of the United States will be served thereby." But his waiver can in no case go so far as to waive a royalty-free license in favor of the United States and foreign countries with which the United States has commitments in this respect.

The Administrator, under Section 305(g) of the Act, can issue licenses under United States patents acquired and can protect his title and rights. He can also require holders under which the Government holds a license to protect such inventions or discoveries. It is not clear whether the Administrator may sue for infringement, require patentees to sue for infringement, or grant licenses subject to royalty payments. In general, under Defense Department procedures, the Government has never asked for a royalty, nor sued anyone for infringement, nor required any holder of patents under which it had rights to sue for infringement.

Proprietary Information

The impact of the Space Act on proprietary rights goes farther than its impact on patent rights. Section 303 of the Act provides that "information obtained or developed by the Administrator in the performance of his functions under the Act shall be made available for public inspection," except security matters and information authorized to be withheld by Federal statute. The word "shall" appears to be one of command.

Section 102(c) provides for interchange of information between Defense Agencies and NASA on a two-way basis. The Defense Agencies are thus required to supply NASA with information, which the latter is required, not authorized, to make available for public inspection.

Under Section 203(a)(3) NASA is required to provide for the "widest practicable and appropriate dissemination of information concerning its activities and the results thereof."

The extent to which proprietary information, such as manufacturing drawings, design information, performance data, and the like, will be disseminated and made public, under the quoted directive, remains to be seen. However, the Act, by its terms, gives the Administrator little leeway. It states that the information "shall be made available for public inspection."

The writer predicts that the provisions of the Act relating to patents will, as actually administered, be found to be certainly no more burdensome than is the Atomic Energy Act, and probably to rather closely parallel the Defense Agency practice. The situation with respect to dissemination of information, on the other hand, may prove to be the sorest point of all, and one which is far more directly of interest to many contractors than are the patent provisions. It is within the personal experience of the writer's firm to have received requests from the Defense Department for full manufacturing information concerning a client's proprietary product, which had been bought off the shelf, on the basis that this was required in order to enable the agency concerned to comply with the Space Act. Because of this there seems to be a current opinion, existing at least in some elements of the Defense Department, that the Space Act requires changes in procedures in the Defense Department.

Contributions Awards

Section 306 of the Act authorizes the Administrator to make monetary awards. He may do this on his own initiative, or upon application of a person or firm, and the amount of the award and the terms upon which the amount may be awarded are left to the discretion of the Administrator. The award must be for a scientific or technical contribution to NASA, which is determined by the Administrator to have significant value in the conduct of aeronautical and space activities. An Inventions and Contributions Board is established, before which hearings may be had by applicants for awards, and has the function of transmitting to the Administrator its recommendations as to the terms of awards to be made.

The factors which the Administrator shall take into account, in making an award, are defined by the Act as follows: 1) the value of the contribution to the United States, 2) the amount of money which the applicant has expended on his own account for the development of the contribution, 3) the amount which the United States has paid toward such expenditure, except that in the case of an officer or employee of the Government salary will not be considered to be such a contribution.

As a consideration for the receipt of such an award, the applicant is required to surrender all claims pertaining to the contribution which he may have against the United States Government or against foreign governments which have suitable agreeements with the United States.

Awards of less than $100,000 appear to be within the sole discretion of the Administrator, but awards exceeding $100,000 require the consequent concurrence of the Congress, which the Congress can convey by failing to object for 30 days after notice to it of intention to grant the award, these 30 days to subsist within a regular session of the Congress.

General Objections

The patent provisions and the proprietary rights of the National Aeronautics and Space Act have been objected to by the Electronic Industry Association. The basic grounds of objection is that the provisions of the Act "may well constitute a first step in the direction leading to the destruction of the American system of patents and trade secrets."

As the patent system is presently established, a long subsisting statute (28USC1498) provides that "whenever an invention described and covered by a patent of the United States is used or manufactured by or for the United States, without license of the owner thereof, or lawful right to use or manufacture the same, the owner's remedy shall be by action against the United States in the Court of Claims for the recovery of his reasonable and entire compensation for such use and manufacture."

The quoted statute in effect provides the United States with a license, not royalty free, under every United States patent. It also provides the patent holder with a procedure for obtaining a reasonable royalty for violation of such patent by the United States, or in its behalf by a contractor, and since the statute provides the sole remedy for the patent holder, no contractor need concern himself with the possibility of an injunction preventing him from fulfilling a Government contract, and the Government in no case can be impeded in procuring equipment, by virtue of any subsisting patent rights. The Government is primarily financially responsible for patent infringement. It can seek reimbursement from a contractor who has infringed without the concurrence of the Government, in certain cases.

The electronics industry recognizes that the Government must adopt procedures which shall assure the Government that it will not be required to pay excessive amounts for equipment which it purchases, and it is only fair that it should not pay a royalty on equipments which it has paid for developing, or toward which it has paid in considerable amount for development. It is for this reason that the electronics industry generally feels that it is reasonable that the Government receive a royalty-free license on inventions made in carrying out research or development work, where the latter is required by a contract with the Government. Industry feels that there can be no objection to the Government filing for and receiving patents based on research which Government has financed, as a protective measure, to prevent subsequent inventors from obtaining patents on the same inventions. However, industry feels that many of the inventions made by industry in the course of performance of research and development contracts, and under which the Government receives free licenses, in fact are made because of a wide area of experience acquired by the contractor without Government support. Where a contractor has acquired a great deal of knowledge and experience, and developed unusual capabilities in certain technical fields, it can be reasonably argued that further inventions made in this field are not fully paid for by the Government merely because the contractor is operating under a Government contract. It is therefore felt that to require contractors to convey to the Government full right, title, and interest to inventions made by them in the course of performance of a Government contract can be, in effect, confiscatory.

Specifically, the industry feels that giving exclusive title to the Government for every invention, made in the performance of any work under any relationship which may exist with NASA, whether or not the Government contributes to the work, or to the invention, and whether or not the invention relates to space activities, presents an excessive tax on industry and does not take account of the realities of the many situations which may arise.

Industry further feels that in providing that the Commissioner of Patents police all patent applications relating to aeronautics and space activities, and in granting him power to refuse to issue any patents thereon unless the applicant therefor clarifies his relationship with respect to NASA, and further requiring certain costly appellate procedures where there is a difference of opinion, involves placing a heavy burden not only on firms having some sort of contractural relation with NASA, but also on firms which do not have such relations. The electronics industry also feels that it is highly inequitable to subject a contractor who enters into a contract with NASA, or who submits a proposal to the Administrator, or who becomes a subcontractor of an NASA contractor, to the danger of having his trade secrets, engineering information, design and manufacturing information, and the like, published and disseminated, regardless of whether or not compensation is afforded therefor.

Industry does not feel that the contributions awards sections of the Act provide an adequate remedy, since in fact they do not provide any legal remedy, but merely provide the Administrator with authority to grant cash awards, within his own discretion.

EMPLOYMENT AGREEMENTS

The question of the conditions under which the Government obtains patent rights, pursuant to the Space Act, deserves some further analysis than has been given. These conditions are established by Section 305(a) of the Act, in two clauses, and the provisions specify that when an invention is made in the performance of any work under any contract of the Administration, and the Administrator makes certain fact determinations, the invention shall belong to the Government. This portion of the Act requires that the invention involved "be made in the performance of any work under any contract of the Administration." However, clause (1) goes on to state that the Administrator can obtain title to the invention whether or not the invention was made during working hours, whether or not the invention was made with a contribution by the Government or the use of Government facilities, equipment, materials, allocated funds, information proprietary to the Government, or services of Government employees during working hours. It is a bit difficult to see how an invention can be made under an NASA contract, without allocation of Government funds, at least, so that these provisions of the Act appear to be contradictory; *i.e.*, the invention must have been made under a contract with NASA, yet the Government need not have contributed in any way to the invention, *i.e.*, need not have paid for the work nor provided facilities or equipment, or provided information. Also, whereas this section states in its initiating phrase that the invention must be made in the performance of work under a contract of the Administration, the following phrases do not make reference to contributions by NASA, but rather to contributions by the Government, which represents a much broader source for contributions. What is probably intended by this disturbing language is that if a contractor enters into a prime contract with NASA, or submits a proposal to NASA, or if a subcontractor submits a proposal to a contractor of NASA, or enters into a sub-contract with a contractor of NASA, and if he thereafter makes (conceives or first reduces to practice) an invention solely with his own funds, facilities, information, and the like, and without any contribution from the Government, that the invention nevertheless belongs to NASA, if NASA can trace some relationship of work done under the NASA contract, to the invention in question. The precise scope of this relationship is not defined, and presumably a relationship of subject matter is intended.

The preceding clause (1) involves a person who made an invention and was employed or assigned to perform research, development, or exploration work, and in such case the invention is required to be "related" to the work he was employed or assigned to perform. Clause (2), however, relates to persons who are not hired to invent or to do research. The inventions of such persons nevertheless belong to NASA if the invention is "related to the contract, or to the work or duties the party was employed to perform, and was made during working hours, or with a contribution from the Government of the sort referred to in clause 1." Such inventions might very well not belong to the employing company under employment contracts in effect, since many companies have employment contracts which exclude from their scope inventions not falling within the scope of the company's business, or providing for retention of certain employee rights under certain conditions, and since many companies do not have invention agreements with employees who are not hired to do research and development, or who are not primarily engineers. Additionally, outside consultants are often treated differently than are full time employees, in these respects.

It therefore appears to be incumbent on contractors who expect to do work for NASA to tighten their employment agreement policies to conform to the provisions of the Act, in respect to employees and consultants. In the first place, employees must have inventions agreements, whether or not the employees are engineers or are assigned to do research and development work, if they might conceivably make an invention. Secondly, the inventions agreement should have no exceptions therein to the classes or types of inventions which the employer can capture, however reasonable these might appear to employer and employee. Any exception can lay the groundwork for a claim by NASA to an invention which the employer has acquired no legal right to convey.

The material presented herein develops those aspects and possible interpretations of the Act which are the most unfavorable from the point of view of industry. We have indicated briefly that the Act may be administered more favorably than is suggested by these adverse interpretations, but we also stress that it is not good Government to pass legislation which is subject to unfavorable interpretation, so that future Administrators thereof, or courts, could at will implement the legislation in a manner which is distinctly unfavorable to those affected by it. At the same time, it is unfair to describe the unfavorable possibilities inherent in the language of the Act without discussing the relaxing regulations under which the Act is currently being implemented. These have been promulgated recently and are found in Vol. 24, No. 212, of the Federal Register, pp. 8788 to 8790.

It will be recalled that the Act gives the Administrator power to waive certain patent rights; *i.e.*, he can take a royalty-free license, and need not take the entire right, title, and interest in any patent, if he believes the interests of the United States will be served thereby.

Thus, the Administrator has established as policy that he may waive his right and assume that such waiver is in the interest of the United States, where "(a) the stimulus of private ownership of patent rights will encourage the development of the invention to the point of practical application earlier than would otherwise be the case, or (b) there are substantial equities justifying the retention of private rights in the invention."

Waiver will not be granted where an invention is primarily adapted for and especially useful in the development and operation of space vehicles and missiles, or where an invention is of basic importance in research in this field, except that in the latter case the Administrator may grant a waiver on such terms as seem to him appropriate.

A *prima facie* case is made out for waiver, 1) where the invention was conceived prior to and independently of entering into an NASA contract, but was first actually reduced to practice in the performance of work under a NASA contract, and the invention is covered by a United States patent issued or application filed prior to the award of the contract, 2) where nonprofit organizations are involved and certain fact situations exist, 3) where the invention has incidental utility for NASA, but substantial promise of commercial utility, 4) where the invention is directed specifically to a line of business of the contractor with respect to which the contractor's previous expenditure of funds has been large in comparison with the amount of funds supplied by NASA.

This portion of the regulation points up the urgency of filing patent applications on inventions before proposals incorporating the inventions therein are submitted to NASA, since such a procedure clearly brings one within the waiver provisions, if the invention is in the proper category.

The regulations proceed to discuss conditions under which a waiver, once granted, can be voided by the Administrator. Waiver granted for a new conception for which a patent application has been filed, or waiver granted because the invention relates to a line of business of the contractor in which it has expended considerable funds, are not subject to voidability. To provide an example, a contractor in the field of telemetry, who conceives a new invention in this field, who applies for a patent thereon, and who thereafter proposes the invention to NASA as a subject for a research and development contract, would be very likely to retain patent rights subject to a free license on behalf of the Government, and such patent rights would not be subject to future voidance.

A waiver granted in respect to an invention which does not enjoy freedom from voidability must be confirmed by certain acts or practices of the patent owner, which must take place on or before the end of the fifth year from the grant of the patent, or the end of the eighth year from the date of acceptance of the waiver, whichever is sooner. The Act, in such case, requires 1) development of the invention to the point of practical application, or 2) making the invention available for licensing either royalty free or at a reasonable royalty basis, or 3) some equitable excuse for failing to fulfil the stated conditions.

Little would be gained by a full discussion of all the details of the regulation in respect to waivers, in an article of this character, but enough has been said to indicate that the actual practice of NASA under the present Act may rather closely parallel, in final and net effect, the Defense Department practice under its present procurement regulations (ASPR).

It further seems reasonable to assume that the entire Administration of the Act, insofar as it relates to proprietary or industrial property will be administered in a manner which appears reasonable and equitable to the Administrator, and, in respect to patents, that the Defense Department practice will be paralleled insofar as the Act permits discretion, and even in respect to those portions of the Act which are not presently dealt with by the regulations. One of the practical reasons for such a conclusion is that NASA cannot hope to obtain the services of sufficient patent attorneys to enable it to capture any large volume of inventions. It must concentrate on the most important inventions. Furthermore, NASA must deal with contractors over long time periods, and in any dealings between any parties, attempts to depart from equitable principles by one of the parties is not good business practice.

The portions of the Act which have not been ameliorated by the regulations, and in respect to which the Administrator appears to have relatively small scope for amelioration, relate to the publication provisions of the Act, and to the capture of technical information from contractors and their subsequent publication. Insofar as the Act is inconsistent with the Copyright statutes, it may well be that the Act supersedes these. It is the belief of the writer, however, that, in due course, regulations will be established in respect to publication, which will be reasonably unobjectionable and will conform more or less with Department of Defense practices, with which industry has learned to live.

It might also be well to stress that the waivers provided for by the regulations are not granted as a matter of right, but rather that the fact situation provides a *prima facie* basis for a waiver. The Administrator is entitled, in each case, to take all the facts into consideration and to make a finding in each case, and, in fact, is required to so do by the Act. It therefore follows that a case may have to be made out by contractors for a waiver in respect to each invention for which a waiver is sought, and therefore that contractors must be very careful to make and keep records consonant with allegations that a waiver in a given case is in the interest of the United States, and that no peculiar equities in favor of the United States and against a waiver have arisen in connection with the conception or the reduction to practice of the invention, or by virtue of the contractor's own commercial practices.

A Comparison of Ion and Plasma Propulsion*

S. W. KASH†

Summary—Several important features and parameters of ion and plasma accelerators for propulsion are compared. Estimates for the thrust of individual ion and plasma acceleration units are given. Impulse measurements for a collinear electrode plasma accelerator are presented. The maximum thrust for an ion gun does not depend on the diameter of the beam, but primarily on the accelerating voltage. Because of the relatively small thrust of the ion accelerator, a greater number of them will be required for a given total amount of thrust. Estimates indicate that the ion accelerator may be somewhat more efficient; however, further experiments are needed to determine the efficiency of both types of accelerators. Beam neutralization is a problem peculiar to the ion accelerator. Considerable research and development may be necessary to provide a satisfactory method for neutralizing the ions. An estimate of the power for neutralization is made. The variation of efficiency with specific impulse is discussed. Further experiments are needed to determine the most efficient ranges of specific impulse for both types of accelerators. Erosion is a serious problem in electrical propulsion; however, for a plasma accelerator it may actually be utilized to provide the propellant material.

INTRODUCTION

IT has been amply demonstrated, that if adequate low-weight power plants were available, it would be advantageous for many space missions to use propellant velocities obtainable only by electromagnetic acceleration. A number of methods for accelerating ionized material have been proposed. Two of these, for which electromagnetic forces provide most of the acceleration, are ion drive and plasma drive.

Basically in all ion-propulsion schemes, ions or charged particles are produced and then electrostatically accelerated. The accelerating electric fields are provided by an arrangement of suitably biased electrodes. Provision must be made for the elimination of electrons of an amount equal to the positive charge and with velocities comparable to that of the positive charge. In plasma-propulsion devices, a neutral plasma discharge is produced and accelerated by rapidly varying intense magnetic fields. The accelerating magnetic fields are often provided by the discharge of large currents through the plasma and associated circuit elements. The large currents are supplied from special low-inductance high-capacitance electrical storage units.

How do ion drive and plasma drive compare? Which is likely to provide the superior propulsion system? These questions must ultimately be answered experimentally. It is quite possible that particular applications will be found for which each system is peculiarly suited. It is worthwhile to explore some of the salient features of ion-drive and plasma-drive accelerators, and to compare several important parameters, including specific thrust, efficiency, specific impulse, neutralization and erosion.

SPECIFIC THRUST

A quantity of importance for electromagnetic propulsion systems is the thrust developed per unit area. This will be referred to as the specific thrust, T_{sp}; it affects the efficiency and mass flow rate and is dependent on the specific impulse. It is apparent that too low a specific thrust will require a large accelerator area and will adversely affect the efficiency and weight of the accelerator.

The specific thrust can be expressed conveniently in terms of Maxwell stresses, that is, in terms of the electric and magnetic fields set up in the accelerators. We can estimate the electric and magnetic fields, respectively, for ion- and plasma-drive devices and compute from these results the maximum T_{sp}.

Ion-Drive Specific Thrust

In an ion-drive device (see Fig. 1), positive ions are produced at an anode and accelerated by an electric field toward a negatively-biased grid or cathode. Figuratively, we may imagine electric field lines emanating from the accelerating cathode and terminating on the ions. The tension along the field lines accelerates the ions backward; simultaneously, the cathode and attached rocket are accelerated forward. The forward thrust on the rocket is obtained from the forces between the cathode and the ions through the electric field tension. Per unit area of accelerator, this tension is $4.42 \times 10^{-7}\ E_c^2$ dynes/cm², where the field at the cathode E_c is expressed in volts/cm. Only field lines ending on ions contribute to the thrust. If the ion current is less than the maximum value, some of the field lines from the cathode will end on the anode and the thrust will be reduced. If E_a represents the field at the anode, T_{sp} will be proportional to $E_c^2 - E_a^2$.

For simplicity, let us assume the usual idealized one-dimensional case, with a separation, d, between the anode and cathode and a voltage drop, V_c, from the anode to cathode. In the absence of any ion current, the field between the anode and cathode will be uniform (equal to V_c/d) and the potential will vary linearly. In this case, $E_a = E_c$ and $T_{sp} = 0$. As the ion current density, j, is increased, some of the field lines will no longer terminate on the anode, but on the intervening ions instead. The potential will vary more slowly near the anode, and hence more rapidly near the cathode; the field at the cathode will be increased and the field at the

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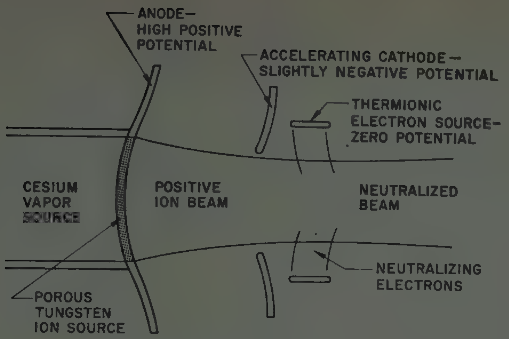
ANODE—
HIGH POSITIVE
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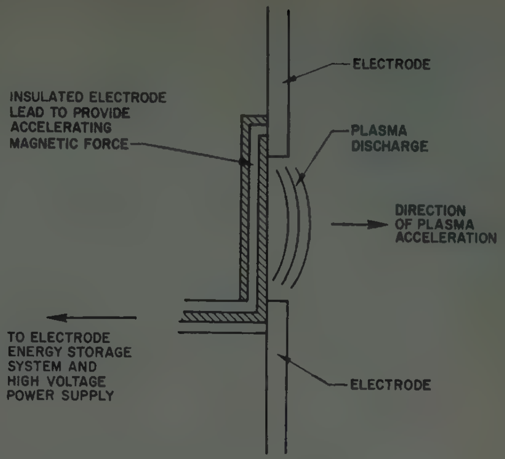
SLIGHTLY NEGATIVE POTENTIAL
THERMIONIC
ELECTRON SOURCE—
ZERO POTENTIAL
CESIUM
VAPOR
SOURCE
POSITIVE
ION BEAM
NEUTRALIZED
BEAM
NEUTRALIZING
ELECTRONS
POROUS
TUNGSTEN
ION SOURCE

Fig. 1—Ion accelerator.

anode decreased. As j increases further, a value will be reached for which all field lines emanating from the cathode terminate on ions and no lines terminate on the anode. The field at the anode will then be zero and no more ions can be accelerated. The current density is said to be space-charge limited and the electric field at the cathode is now at a maximum.

The space-charge-limited case ($j=j_s$) is readily analyzed. First, equate the electric tension $4.42\times10^{-7}\ E^2$ with the momentum flux nmv, where m and v are the mass and velocity, respectively, of the ions, and n is the number of ions per second crossing a square centimeter. (The current density is simply n times the charge e per ion, or eT_{sp}/mv_c.) Now, express E and v in terms of the potential drop V and the distance x and integrate. The resulting familiar expression shows that the potential varies as the distance to the 4/3 power. Accordingly, the field at the cathode, which is now at a maximum, is 4/3 the zero current value and the maximum T_{sp} is $7.86\times10^{-7}\ (V_c/d)^2$ dynes/cm² or $16.4\times10^{-10}\ (V_c/d)^2$ pounds/foot².

It should be emphasized that this result is entirely independent of the mass or charge of the particles. As an example, for an applied voltage of 10,000 volts and an interelectrode spacing of one centimeter the specific thrust is at most only 0.16 pound/foot². For singly ionized cesium ions, this would provide a velocity of 1.2×10^7 cm/second and a maximum current density of 4.7 ma/cm².


ELECTRODE
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ENERGY STORAGE
SYSTEM AND
HIGH VOLTAGE
POWER SUPPLY
ELECTRODE

Fig. 2—Collinear electrode plasma accelerator.

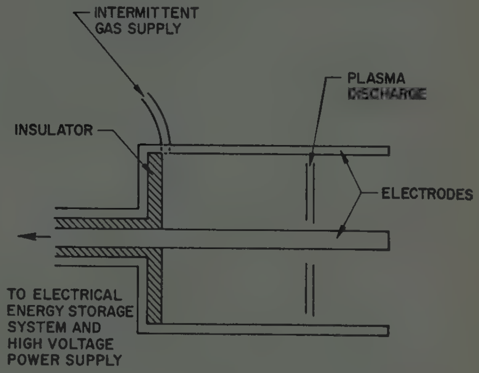
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Fig. 3—Coaxial electrode plasma accelerator.

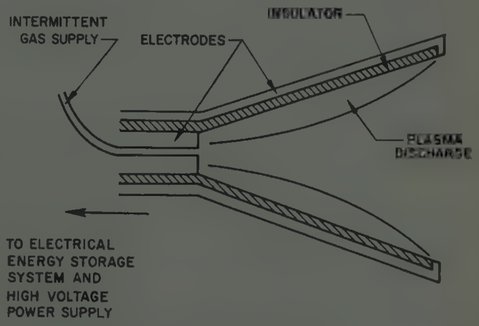
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Fig. 4—Plasma pinch accelerator.

Plasma-Drive Specific Thrust

In comparison with ion propulsion, a greater complexity of devices is available for plasma propulsion. Figs. 2–4 show three different electrode configurations capable of accelerating plasma. In each of these, a plasma introduced into the discharge region or produced by the discharge, is accelerated by the interaction of the current in the plasma and the current in the fixed part of the discharge circuit. Generally, the current elements in the parts of the circuit nearest the plasma contribute the most to the acceleration.

In the collinear electrode plasma accelerator (see Fig. 2), the principal force arises from the repulsion between the current element in the plasma and the antiparallel current element in the adjacent portion of the circuit. In the coaxial electrode system (see Fig. 3), the acceleration is provided by the repulsive interaction between the disk-sheet plasma current and the current in the central electrode. For the pinch accelerator, shown in Fig. 4, the accelerating force arises from the attractive self-interaction of the axially symmetric current sheet. Here the radial component of the current in the plasma contributes directly to the thrust; the axial component contributes indirectly through plasma compression and heating, much as in chemical propulsion.

For each system described, pulsed action is presumed and no steady magnetic fields are applied. In operation, large currents are obtained from the discharge of low-inductance high-energy capacitors. With careful design, the rate of rise of the magnetic field and the plasma conductivity can be sufficiently large so that there is little penetration of the magnetic field into the plasma during the acceleration. In this case the magnetic field in effect acts as a piston, and, for a field of H gauss, the magnetic pressure or T_{sp} is 0.04 H^2 dynes/cm². For example, for a field of 10,000 gauss, the specific thrust would be approximately 4×10^6 dynes/cm², or 8.3×10^3 pounds/foot².

Limitation on Ion Specific Thrust

As shown above, the specific thrust of an ion accelerator depends only upon the electric field intensity at the accelerating electrode, and is largely determined by the voltage and spacing between the cathode and the anode. The voltage itself is determined by the specific impulse and the mass of the ion. For example, with singly ionized cesium ions and an I_{sp} of 10^4 seconds, a voltage of about 7000 volts is required. Thus, to maximize T_{sp} it is necessary to minimize the interelectrode spacing.

Ion optics sets a limit on how small the interelectrode spacing d can be. Let D be the beam or cathode aperture diameter; we presume that to reduce erosion the aperture does not contain grid wires. As is the case for electron guns,[1] when d/D is less than unity, the field pattern within the ion beam should be different from that in the absence of the cathode aperture, and the average current density should be less than the space-charge-limited value based upon parallel-plane electrode theory. Furthermore, when d is less than D, there will be increased divergence of the beam. Thus, if a low-divergence high-intensity beam is to be obtained from the whole anode area, the interelectrode spacing should be larger than the beam diameter. With an accelerating potential of 10^4 volts, corresponding to an I_{sp} of 12,000 seconds for singly ionized cesium, and with a beam diameter and interelectrode spacing of, say, one centimeter, a maximum specific thrust less than 100 dynes/cm² is to be expected.

Argon-ion beam experiments[2] support the ion-optics considerations presented above. In these experiments, the equivalent perveance was measured for a number of circular-aperture arrangements at three different interelectrode spacings d. Table I lists microperveance values μ for the arrangement near maximum at all three spacings. The diameter of the cathode aperture is 1.0 inch, and the d values are based upon the beam profiles given by Harrison.[3] The last column presents the product $\mu(d/D)^2$ and shows the departure from parallel-plane electrode theory at the smallest d/D value. For $d/D = 1.5$, μ is close to the value predicted for parallel-plane electrodes. Although at $d/D = 0.5$ the current density is considerably less than the parallel-plane space-charge-limited value, the perveance (total current/voltage$^{3/2}$) itself is actually larger and is probably near maximum. It appears preferable, nevertheless, to use a larger spacing such as $d = D$, in order to reduce erosion.

TABLE I

VARIATION OF MICROPERVEANCE WITH d/D RATIO

Case	d/D	μ	$\mu(d/D)^2$
c	0.5	5.9	1.5
h	1.0	2.3	2.3
k	1.5	1.1	2.5

Consideration may be given to the use of smaller diameter beams with correspondingly smaller interelectrode spacings. For example, for a 3-mm diameter beam with a 3-mm interelectrode spacing and an accelerating voltage of 10 kv, the maximum specific thrust would be 873 dynes/cm², corresponding to a maximum current density of 52 ma/cm². It is interesting to note that for a given accelerating voltage and a fixed d/D ratio, the total thrust remains constant and independent of beam diameter. This results from the fact that the increase in T_{sp} is exactly compensated for by the decrease in beam cross section. Thus, the total number of ion guns for a given vehicle thrust will not vary greatly with ion-beam diameter. For an accelerating potential of 10 kv and with $d = D$, the maximum thrust per ion gun is 62 dynes, and about 7200 of them would be required to produce one pound of thrust.

To reduce radiation losses, smaller diameter beams are preferable. There are several factors, however, which provide a lower limit on the beam diameter or interelectrode spacing. For one, smaller diameters require larger current densities to be drawn from the ion source, which in turn requires a somewhat higher anode temperature,[4] and may also increase the erosion of critical electrode surfaces. Again, for a given voltage there is a minimum electrode separation set by the electrical-breakdown voltage gradient.

Limitation on Plasma Specific Thrust

For a plasma accelerator, the specific thrust depends upon the magnetic field intensity. In turn, the magnetic field is determined by the discharge current and geometry. If I is the current (in amperes) and d is a characteristic electrode dimension (in centimeters), H is of the order of $I/5d$ and T_{sp} is of the order of 10^{-3} $(I/d)^2$ dynes/cm². Now d can be of the order of a centimeter or less and I of the order of 10^5 amperes; hence, T_{sp} can be of the order of 10^7 dynes/cm².

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[1] J. R. Pierce, "Electron guns," in "Theory and Design of Electron Beams," D. Van Nostrand Co., Inc., New York, N. Y., ch. 10; 1949.

[2] E. R. Harrison, "Investigation of the perveance and beam profiles of an aperture disc emission system," *J. Appl. Phys.*, vol. 29, pp. 909–913; June, 1958.

[3] *Ibid.*, p. 911, Fig. 5.

[4] J. B. Taylor and I. Langmuir, "Vapor pressure of cesium by the positive ion method," *Phys. Rev.*, vol. 51, pp. 752–760; 1937.

In practice, such a high specific thrust cannot be achieved. Whereas the electric fields in an ion device act continuously, the magnetic fields in the plasma devices described act only intermittently. If we assume that the field is effective only 0.1 per cent of the time, the plasma-drive specific thrust is of the order of 10^4 dynes/cm^2. This is well over one order of magnitude greater than the maximum T_{sp} of an ion-drive device.

For a given peak current, the specific thrust will decrease with increasing aperture of the plasma accelerator, somewhat as it does with the ion accelerator. However, the plasma accelerator is not space-charge-limited, and a greater amount of power can be utilized in a single unit to accelerate a greater amount of propellant. It is possible to build a plasma accelerator with an effective area of between 10 and 100 square centimeters and still maintain a high specific thrust. Thus, a single plasma accelerator may be capable of providing a thrust of 0.1 pound or more.

Measurements on the Specific Thrust of a Plasma Accelerator

Experiments have been performed with a collinear electrode plasma accelerator, and measurements of impulse and efficiency have been made.[5] The specific thrusts obtained are in accord with the values estimated. In essence, the device (see Fig. 2) consists of a pair of collinear discharge electrodes connected to a low-inductance high-energy capacitor. The leads between the capacitor and the electrodes are arranged to keep the inductance of the discharge circuit as low as possible and also to provide the accelerating force for the plasma.

In operation, the region around the electrodes was evacuated and the capacitor was charged to a high voltage. A discharge was initiated by the introduction of plasma between the electrodes. For one set of measurements, the plasma for the discharge was produced by exploding a fine wire mounted between the electrodes and at right angles to the plane of the discharge circuit. In a second set of measurements, the discharge was initiated by the introduction of a minute amount of gas between the electrodes, with the major portion of the plasma being supplied by the electrodes themselves.

With the exploding-wire system a number of firings were made using different materials and different diameter wires. The vaporization of the fine wire was accomplished by discharging a small capacitor through it. Impulse measurements obtained are shown in the second column of Table II. The mass of the wire vaporized is shown in the third column, and the effective specific impulse determined from the impulse and wire mass is shown in the fourth column. The final column presents the efficiency obtained by comparing the computed kinetic energy of the ejected plasma with the initial energy stored in the main capacitor. As can be seen, the smaller-size wires yield the greater specific impulses and greater efficiency. Higher velocities and efficiencies than those presented should be possible by further improvement of the system.

For the expendable electrode system a small piston was used to admit a minute amount of triggering argon gas through a hole in the upper electrode. The impulses obtained were not sensitive to the amount of gas admitted, and as little as six micrograms of argon was adequate to trigger the discharge. The major portion of the propellant was obtained by erosion of the electrodes.

TABLE II

RESULTS OF IMPULSE MEASUREMENTS FOR THE EXPLODING-WIRE SYSTEM

Wire size	Measured impulse (dyne·seconds)	Mass of wire (mg)	Effective specific impulse (seconds)	Efficiency (per cent)
1 mil platinum	1380	0.90	1530	21
5 mil nichrome	1620	6.4	260	4
1 mil tungsten	1800	0.83	2210	39
2 mil tungsten	1750	3.3	540	9
5 mil nichrome	1750	6.4	279	5

TABLE III

VARIATION OF IMPULSE WITH VOLTAGE FOR THE EXPENDABLE ELECTRODE SYSTEM

Capacitor voltage V (kv)	Measured impulse I (dyne·seconds)	I/V^2 (arbitrary units)	CV^2/I (cm/second)
15	160	0.711	15.5×10^6
20	320	0.800	13.8
25	416	0.655	16.8
32	832	0.812	13.5
35	960	0.783	14.0

The impulse measurements for this system are shown in Table III. Analysis shows that both the impulse and the mass eroded from the electrodes should be proportional to the square of the initial voltage. The proportionality between the impulse and the square of the voltage is experimentally demonstrated by the results shown in the third column of Table III. A further consequence of the analysis is that both the specific impulse and the efficiency should be independent of the initial voltage. The specific impulse can be changed by varying the composition and/or the spacing of the electrodes.

Column four presents the effective velocity if the efficiency were one hundred per cent. With an assumed efficiency of 40 per cent the effective velocity computed from these data is about 6×10^6 cm/second. For this velocity the propellant mass ejected for a 35-kv firing is computed to be about 160 micrograms. With less than six micrograms of gas needed to trigger the discharge, over 96 per cent of the propellant is thus provided by the electrodes.

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[5] S. W. Kash and W. L. Starr, "Experimental Results with a Collinear Electrode Accelerator and a Comparison with Ion Accelerators," presented at American Rocket Society Meeting, Washington, D. C., November, 1959, paper no. 1008–59.

The data obtained with the collinear electrode system agree with the specific-thrust estimates made earlier for plasma accelerators. The end-on area of this device is of the order of 10 cm^2 and the firing time is of the order of 10^{-5} seconds. With about 10^3 dyne·seconds for the total impulse, the specific thrust is 10^7 dynes/cm^2. For a firing rate of one per second, the collinear electrode device produces between one and two grams of thrust. Only several hundred firings per second are necessary to produce a pound of thrust. This might be accomplished by a single device operating at this rate or by several similar devices operating at lower repetition rates.

Suggested Ways for Improving the Specific Thrust for Ion Drive

Two ways have been proposed for increasing T_{sp} for an ion accelerator. In the first, heavy-charged particles are suggested in lieu of ions. This would require the use of higher voltages to obtain the necessary specific impulse and would presumably yield more mass for a given current density and hence more thrust. The second way suggested is to use a double-grid accelerate-decelerate system to overcome the space-charge limitation on the current density.

In both of these schemes, higher voltages would be used to obtain the larger gradients necessary for an increase in T_{sp}. Unfortunately, if both the voltage and the voltage gradient are increased, electrical breakdown will ensue. Experiments have shown[6] that for vacuum insulation the maximum voltage and maximum voltage-gradient are interrelated. The results of these high-voltage breakdown experiments are shown in Fig. 5. Large gradients of the order of several million volts per centimeter can be sustained for voltages below 20 or 30 kv. Conversely, large voltages of the order of 1 million volts can be sustained only for small gradients, about 10^5 v/cm or less. Over a fairly wide range of voltage breakdown, the product of the voltage and voltage gradient is sensibly constant.

The breakdown data apply to cold, well-polished and outgassed metal surfaces, and result from a cascading of cathode electron emission and subsequent bombardment at the anode, and from ion and photon emission and subsequent bombardment of the cathode. Much smaller gradients and voltages can be sustained, with heated irregular surfaces, such as a porous tungsten anode and a partially-eroded cathode, and, of course, the deliberate presence of ions. Voltage gradients averaged over the plane of the cathode will be well below 10^5 volts/cm for applied voltages of the order of 10 kv.

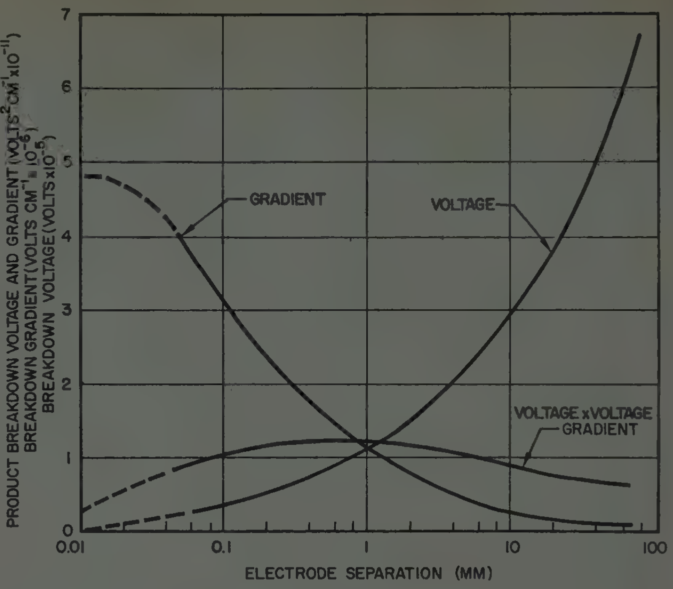

Fig. 5—Vacuum breakdown voltage between a one-inch stainless-steel ball and a two-inch stainless-steel disk. (After Trump.[6]) The lower curve has been added to illustrate the approximate constancy of the product of the breakdown voltage and the gradient.

If a higher voltage can be had only at the expense of a lowered voltage gradient, it is clear that little if any increase in T_{sp} can be obtained with a double-grid accelerate-decelerate system. In the heavy charged-particle accelerator, T_{sp} will actually be decreased; for even if efficient means were available to provide the micron or submicron particles with the maximum charge possible, accelerating voltages of the order of a million volts or more would be required to obtain a satisfactory specific impulse. This would entail a serious reduction in the voltage gradient and thus an even greater reduction in T_{sp}.

NEUTRALIZATION

For the continued operation of an electrical propulsion system it is necessary to maintain charge neutrality on the vehicle. This requires the ejection of as many electrons as positive ions. Furthermore, the electrons must leave the space vehicle with about the same speed as that of the ions. For a plasma-drive device these conditions are automatically satisfied. On the other hand, for an ion-drive device special provision must be made to supply the neutralizing electrons.

It is not obvious that large-area ion beams can be efficiently neutralized. Neither an analytical nor an experimental demonstration to this effect has yet been presented. The major difficulty arises from the large disparity between the ion and electron masses. Experiments have shown that the current in a pencil-thick ion beam can be neutralized by a peripheral thermionic electron source.[7] However, these experiments have not, as yet, shown that charge neutrality was established downstream in the ion beam. This requires measurements to show that the ions and electrons have achieved the same axial velocity. It is possible, and indeed quite likely, that the neutralizing electron currents obtained in the experiment constitute a sparser but faster electron beam.

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[6] J. G. Trump, "Insulation strength of high-pressure gases and of vacuum," in "Dielectric Materials and Applications," John Wiley and Sons, Inc., New York, N. Y., pp. 147–156; 1954.

[7] R. C. Speiser and C. R. Dulgeroff, "Cesium Ion Motor Research," presented at Air Force Office of Specific Research Second Symposium on Advanced Propulsion Concepts, October, 1959; to be published.

It has been pointed out that if an ion beam is surrounded by an annular thermionic electron source, electrons can be drawn into the beam with large radial velocities and with average energies of about 2000 volts.[8] This, of course, represents a serious loss of energy and would occur at the expense of ion-beam energy. However, it has also been recognized that the presence of other electrons would reduce the electrical potential within the ion beam and reduce the energy loss. One estimate reduces the acquired electron energy to about half,[9] which still represents a considerable energy loss. Since the potential on the axis of a long cylinder of charge depends on the charge per unit length, the potential on the axis will also be fairly independent of beam diameter.

A complete analysis of the neutralization problem has apparently not yet been made. It is quite possible that if the ion beam were surrounded by an extremely copious supply of electrons, the neutralization within the beam would be sufficiently complete and the electrons would acquire only a few volts of energy upon being drawn into the beam. There is a practical difficulty involved here in that the electron source is also space-charge limited and a large potential gradient and/or large electron source area is needed to get a sufficient electron current. Fortunately, the electron mass is small, so that more modest gradients than needed for an equivalent ion-current density will suffice. If the electron-emitting area is comparable to the ion-source area, and if its distance to the center of the ion beam is comparable to the ion-interelectrode distance, then a potential of only a few hundred volts will be needed to obtain the necessary electron current.

We may expect, then, some balance to be achieved by which the potential at the center of the ion beam relative to the electron-emitting surface is slightly above that necessary to provide the neutralizing electrons. There will still be a substantial energy loss in the form of electron kinetic energy. It should be possible, in principle at least, to reduce this further by the use of special electron accelerating surfaces at the electron source. Modest voltages below 100 volts would probably be sufficient. Thus, the loss to radial electron kinetic energy might be reduced, although at the expense of greater complexity of design and operation.

EFFICIENCY

The kinetic power and mass-flow rate per unit area are readily expressed in terms of the specific impulse and specific thrust. With I_{sp} in seconds and T_{sp} in dynes/cm², these are, respectively,

$$P_{sp} = 5 \times 10^{-8} g I_{sp} T_{sp} \text{ watts/cm}^2$$

$$M_{sp} = T_{sp}/g I_{sp} \text{ gm/cm}^2 \text{ seconds.}$$

For a cesium-ion accelerator with an accelerating voltage of 10 kv and an anode to grid spacing of only one centimeter, $I_{sp} = 1.2 \times 10^4$ seconds, a maximum $T_{sp} = 78$ dynes/cm², a maximum $P_{sp} = 47$ watts/cm², and a maximum $M_{sp} = 6.5$ micrograms/cm² seconds. For a plasma accelerator with a comparable I_{sp} and a magnetic field of only 10^4 gauss, and operating only 0.1 per cent of the time, the maximum $T_{sp} = 4000$ dynes/cm², the maximum $P_{sp} = 2350$ watts/cm², and the maximum $M_{sp} = 340$ gm/cm² seconds.

P_{sp} represents the power usefully converted. To obtain the efficiency one must take into account all other power required or lost. For the ion drive this includes power radiated by the anode and surrounding structure, power for neutralizing electrons, power to vaporize and ionize the ions, and power lost because of sputtering. For the plasma drive this includes the power to produce the plasma, power radiated by the plasma, power lost in the electrical circuits and so forth. Account must also be taken with each system for any loss of unaccelerated propellant material.

One of the largest power losses for an ion accelerator results from thermal radiation from the anode ion source, the thermionic electron source and nearby heated exposed surfaces. As pointed out earlier, reducing the ion-beam aperture need not reduce the thrust, but it certainly can reduce the amount of exposed anode radiating surface. This reduction will be offset somewhat by the need for a higher anode-surface temperature to provide the increased current density. Consider a 5-mm diameter cesium-ion gun operated at 10 kv. This would require an anode current density of 19 ma/cm² and would give an average voltage gradient of 20 kv/cm, both realistic values. The energy flux in the ion beam would be about 37 watts. For the current density indicated, an anode temperature of about 1400°K would be required. This would radiate about 20 watts/cm² for a total of about four watts. We should expect that the surrounding thermionic electron emitter will contribute at least an equal amount to the radiation flux. Thus, the radiation loss will be at least 20 per cent that of the kinetic energy in the ion beam. This must be considered a lower limit. If one uses larger diameter beams, the same thrust can be obtained, but with a considerably larger radiating anode surface. On the other hand, we may expect conduction from an array of small high-temperature anodes to bring some of the surrounding structure to a high temperature, so that there may well be significant radiation losses from surfaces other than the ion anode and electron cathode.

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[8] E. Stuhlinger and R. Seitz, "Some problems in ionic propulsion systems," IRE TRANS. ON MILITARY ELECTRONICS, vol. MIL-3, pp. 27–33; April, 1959.

[9] G. G. Cloutier and T. W. Johnston, "Charge Neutralization in an Ionic Rocket," RCA Victor Co., Montreal, Que., Canada, Res. Rept. No. 7-811-1; August, 1959.

If we suppose that the neutralizing electron source is a simple emitting surface with no special accelerating electrodes, then the electrons will acquire radial kinetic energy of as much as 1000 ev per electron. If we assume further one electron per ion, the neutralizing electrons will carry away an amount of energy about ten per cent that of the ion beam energy. This energy will not, of course, contribute significantly to the thrust.

Erosion losses are difficult to estimate. The amount of erosion can no doubt be reduced by better ion optics; but this most likely requires a reduction of the total beam current with a corresponding reduction of beam power and over-all efficiency. Let us assume a modest erosion loss of, say, two per cent; that is, we assume that only two per cent of our ions strike the cathode. These collisions will, by sputtering, release an amount of material between five and ten times that of the incident ions. This material will not contribute appreciably to the thrust. We should also add to this one to two per cent of the mass efflux for the loss of un-ionized cesium vapor. The loss of unaccelerated material in effect reduces the specific impulse or thrust of the propellant by at least ten per cent. To compensate for this, the velocity of the ions can be increased one or two per cent, which would require an additional power increment of two to four per cent. Again, this may be an optimistic estimate. The sputtering process will also release electrons, most of which will be drawn to the anode, where they will give up their acquired energy. This can contribute to the anode heating and need not be considered an energy loss.

Some additional power will also be required to vaporize and ionize the cesium, but this should be quite small, less than one per cent. Collecting the more favorable estimates of power loss caused by radiation, neutralization, erosion, and ion production, one obtains an efficiency of between 70 and 75 per cent. A shortcoming of this estimate on over-all efficiency is that to a large extent it presupposes that the various losses cited can be independently minimized. In general, this may not be possible. For example, the electrode configuration that minimizes erosion may also yield a reduced thrust. As a second example, higher current densities may be accompanied by faster destruction of the anode and by greater loss of un-ionized propellant.

In considering the efficiency of a plasma-drive device, we can conveniently divide the problem as follows. How efficiently can the electrical energy be delivered from the storage unit to the plasma acceleration unit, and what fraction of the energy delivered goes to produce useful kinetic energy? Because of the transient nature of the system, an exact analysis is extremely difficult. To transfer energy efficiently from the capacitance storage unit to the plasma acceleration unit, the characteristic impedance of the storage unit and the resistive impedance of the plasma circuit should be comparable. Otherwise, a substantial fraction of the energy will be reflected back into the storage unit and the system will oscillate or ring. Fortunately, an exact matching of impedances is not necessary. Even for a 50 per cent mismatch, about 90 per cent of the energy will be transferred during the first cycle, and additional energy will be transferred in subsequent rings of the system.

All fixed resistance and inductance of the system can be lumped with the capacitor storage parameters. The resistance of the plasma itself is at first largely ohmic in nature and provides the heating necessary to produce and ionize the plasma. This resistance quickly falls with rising plasma temperature. During the acceleration of the plasma, the resistive component results from the rapidly changing inductance of the plasma discharge portion of the circuit. Because the discharge inductance is very small, the impedance ratio in present laboratory systems is low (between 0.01 and 0.1) and there is considerable ringing of the system. By proper design of the total system it should be possible to increase the impedance ratio to over 0.1. Even with a ratio of only 0.1, about 30 per cent of the energy is transferred during each pulse, and about five pulses are required to transfer over 80 per cent of the energy.

Not all of this will go into useful kinetic energy of the plasma. Some of it is required to produce the plasma, and a good portion of it will contribute to heating the plasma and circuit elements. The fraction converted to plasma kinetic energy will depend upon the design of the electrodes and storage unit, and upon the amount and type of plasma. With proper design, well over 70 per cent should go into plasma kinetic energy. Accordingly, we may expect that the over-all conversion of stored electrical energy to plasma kinetic energy should be 50 per cent or better. Indeed, in the plasma thrust experiments mentioned earlier, where the design was far from optimum, efficiencies of over 30 per cent have been obtained.

SPECIFIC IMPULSE AND EROSION

For the collinear electrode plasma accelerator, both analysis and experiment indicate that the efficiency and the specific impulse are not affected by the specific thrust. How the efficiency varies with I_{sp} is not certain, but it appears that it should still increase somewhat with increasing I_{sp}. There undoubtedly exists a range of I_{sp} for which the efficiency is near maximum.

In ion drive, the need for a high-temperature ionizing surface limits the range of I_{sp}. At large I_{sp}, because of the voltage-breakdown limitations discussed earlier, an increase in the voltage necessitates roughly an inverse decrease in the voltage gradient. As a result, the ratio of the power in the ion beam to the power radiated varies roughly as the inverse cube of I_{sp}. On the other hand, at low I_{sp}, since the anode-to-cathode spacing cannot be indefinitely reduced, the voltage gradient is proportional to the voltage. Hence, the ratio of beam power to radiated power varies approximately as I_{sp}^5. Thus, the efficiency of an ion accelerator decreases rapidly at both high and low I_{sp}.

A comparable situation for plasma drive would exist if, for example, power were required for a steady-state magnetic field. This, however, can be avoided by a suitable design of plasma accelerator. Thus, the efficiency of a plasma-drive device need not fall with operation at lower specific thrust. Further experiments are needed to determine the efficient ranges of I_{sp} for both ion and plasma devices. It may be added that, because of its intermittent type of operation, a plasma-propulsion unit should operate efficiently over a wide range of thrust.

Erosion presents a particularly serious problem for electrical propulsion devices. In ion drive, erosion can seriously impair the efficiency and reliability of the accelerator.[10] Because of erosion, an accelerating cathode in the form of a grid mesh does not appear very suitable. Good focusing geometry may reduce the erosion to a small amount. However, this may entail an increase in the anode-to-cathode distance with a corresponding decrease in the specific thrust and efficiency.

Erosion need not be as serious a problem for plasma drive. In fact, as shown experimentally with the collinear electrode plasma accelerator, erosion can be employed to provide the major portion, if not all, of the propellant material.

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[10] S. Naiditch, "Experimental Ion Sources for Propulsion," presented at American Rocket Society Meeting, San Diego, Calif., June, 1959, paper no. 833-59.

Comparison of Chemical and Electric Propulsion Systems for Interplanetary Travel*

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Summary—The basic mission parameters which are required for the evaluation of engine performance for interplanetary flights are defined. The engine parameters are also defined, and the relation between these parameters is formulated. The possibility of achieving much larger payload fractions by the use of electric propulsion systems, as opposed to the use of chemical propulsion systems, is indicated. Methods of calculating impulsive orbit transfers between circular orbits, escape and entry from satellite orbits using low thrust, and optimized powered transfer between heliocentric orbits are given in the Appendix.

INTRODUCTION

A FUNDAMENTAL problem in present-day space flight is attainment of adequate payload. This problem originates in the limitations of the chemical propulsion systems utilized. In search of methods to circumvent such limitations electric propulsion techniques are presently the object of considerable interest. Historically, improvement in space vehicle performance has come about through increase in propellant specific impulse. Since electric acceleration techniques are known to produce extremely high-particle velocities in related sciences, the extension of such electric techniques into the propulsion field is quite natural. The availability of high specific impulse is not an unmitigated attribute, however, since in electric propulsion techniques the energy required to accelerate the propellant is no longer contained in the propellant itself. Such energy is provided by a solar or nuclear electric supply which, with today's technology, has considerable weight. This weight must be considered in evaluating the gains provided by use of high specific-impulse electric propulsion systems.

The object of this paper is to set forth the basic parameters and to evaluate the capabilities of electric propulsion systems. Comparisons will be drawn with capabilities of conventional chemical systems. This evaluation can best be carried out by considering specific missions. The missions selected for this analysis are the Earth-Mars and the Earth-Venus missions. The engines, as characterized by specific mass, exhaust velocity, mass flow rate, and thrust, will be discussed in the following section. The definition of the mission and its mathematical formulation will be given in the third section. The comparison and evaluation are in the last section. The details of the calculations are given in the Appendix.

The results of such an analysis are important in two ways. First, the mission capabilities of electric propulsion systems being studied can be determined, and second, some guidance in the selection of desirable directions for research can be obtained.

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PROPULSION SYSTEM PERFORMANCE PARAMETERS

The rocket engine operates by ejecting mass. In electric systems, charged particles are accelerated by electric, or electric and magnetic, fields and expelled. The substance expelled is called the propellant. In chemical systems, the expellant is energized by a thermochemical reaction which uses the propellant. If M is the rocket mass, $\dot{M}$ is the time rate of change of rocket mass, $\overline{C}$ is the effective propellant or exhaust velocity relative to the rocket, and $\overline{F}$ is the thrust force due to the rocket engine, then

$$\overline{F} = \dot{M}\overline{C}. \tag{1}$$

The thrust force per unit mass of the rocket, $\bar{f}$, is defined by

$$M\bar{f} = \overline{F}. \tag{2}$$

If $\bar{a}$ is the acceleration vector and if all forces except the thrust force are ignored, it follows from (1) and (2) that

$$M\bar{a} = \dot{M}\overline{C}. \tag{3}$$

It should be noted that $\bar{a} = \bar{f}$ if no other forces are acting. This is the well known rocket equation. For the case where the velocity, direction and rate of propellant flow are constant, (3) can be written as a scalar equation, and integrated immediately to get

$$\Delta v = v_1 - v_0 = C \log \frac{M_0}{M_1}, \tag{4}$$

where the subscripts 0 and 1 refer to the initial and terminal values respectively, v is the velocity, and Δv is called the characteristic velocity. This may be written in terms of the specific impulse I_{sp} as

$$\Delta v = g_0 I_{sp} \log \frac{M_0}{M_1}, \tag{5}$$

where M_0/M_1 is called the mass ratio, and the specific impulse is defined by

$$I_{sp} = \frac{F}{\dot{w}} = \frac{C}{g_0}. \tag{6}$$

Two systems of units are commonly used for defining the specific impulse. In one, F is in pounds force and $\dot{w}$ is in pounds mass per second, and the specific impulse is in pounds force per pounds mass per second. In the other system which will be used here, F is in pounds force and $\dot{w}$ is in pounds weight per second. Although both systems give the same numerical value for the specific impulse, in the second system, the specific impulse is in seconds.

An approximate indication of the terms on which specific impulse can be exchanged for mass flow rate for a given characteristic velocity can be obtained from (5). In general, a slight change in the specific impulse can be balanced only by a large change in the mass ratio. For electric propulsion systems, the specific impulse ranges approximately between 1000 and 20,000 seconds for practical purposes, while for chemical systems, the specific impulse is under 500 seconds. Thus, for a given Δv the electric system requires considerably less propellant mass than a chemical system. Eq. (5) also makes evident that when extremely large Δv's are required, typical of ambitious missions, the use of very high specific-impulse electric systems is the only way by which impossibly large mass ratios can be avoided.

However, the conclusion that the payload is larger for an electric propulsion system than for a chemical propulsion system is not necessarily valid, because for an electric system the mass M_w of the power source and acceleration system is an important factor. The parameter used for this factor is the specific mass which is defined by

$$\alpha = M_w/P, \tag{7}$$

where P is the jet power:

$$P = 1/2(-\dot{M})C^2. \tag{8}$$

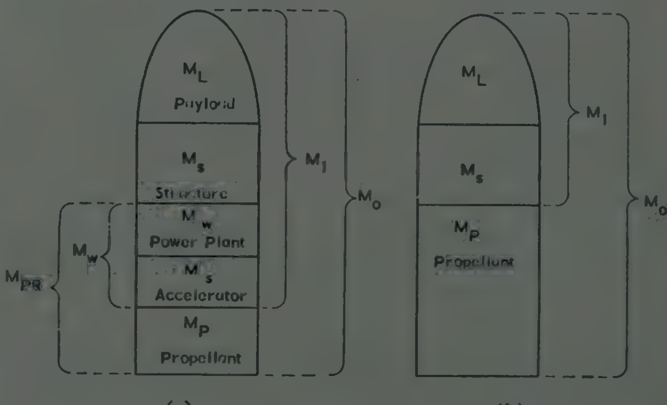
The values of α considered in the literature range from five to fifty kilograms per kilowatt.

A more detailed analysis of a propulsion system is made in the following way. The vehicle is regarded as consisting of two parts. One part is the structure of mass M_s and payload of mass M_L which are independent of the propulsion system. The other part is the propulsion system of mass M_{PR} which consists initially of a power plant of mass M_w', a propellant mass M_p, and a mass M_s' which accounts for the remainder of the propulsion system, including the accelerating device. This last mass is assumed to be proportional to the maximum available thrust and is described by the specific mass

$$\beta = M_s'/F_{\max}. \tag{9}$$

These masses are illustrated in Fig. 1. Also, for this detailed analysis, α' is defined only with reference to the power source; *i.e.*,

$$\alpha' = M_w'/P_e, \tag{10}$$



(a) (b)

Fig. 1—Schematic of vehicle mass distribution. (a) Electric propulsion system. (b) Chemical propulsion system.

where P_e is electrical power output. In addition, the efficiency factors ξ and η are used to bridge the gap between the designer of the electric propulsion system and the mission analyst. The parameter ξ is the proportion of propellant utilized, and the parameter η is the ratio of the directed kinetic energy to the total electric energy input. In these parameters, the total mass of the propulsion system M_{PR} can be represented as

$$M_{PR} = (\alpha' P/\eta) + \frac{1}{\xi}\int_0^T (-\dot{M})dt + \beta F_{\max}, \qquad (11)$$

where P is the jet power and T is the duration of the jet operation. If the rate of mass flow is constant,

$$M_{PR} = (\alpha' P/\eta) - (\dot{M}T/\xi) + \beta F_{\max}. \qquad (12)$$

For the purposes of this paper, it will be assumed that the first and third terms are proportional or that $\beta F_{\max}$ is negligible with respect to (αP). Thus, by (7) the equations corresponding to (11) and (12) are

$$M_{PR} = \alpha P + \int_0^T [-\dot{M}]dt, \qquad (13)$$

and for a constant rate of mass flow

$$M_{PR} = \alpha P + [-\dot{M}]T, \qquad (14)$$

where it is assumed that ξ is set equal to 1.0, its maximum possible value. It should be noted at this point that the terms η, β and ξ which, for convenience in this paper, have been lumped together or assumed to be unity, are actually prime design parameters of electric propulsion devices. The degree to which these parameters are minimized or maximized, as appropriate, is a measure of the excellence of the particular thrust unit considered.

Two additional parameters which describe the vehicle are the payload ratio (M_L/M_0) and the structure factor K. If M_0 is the total initial mass of the vehicle, then K is defined by

$$K = \frac{M_s}{M_0 - M_L}\cdot \qquad (15)$$

For the purposes of this paper, the value of K will be assumed to be 0.05. In terms of the mission and engine parameters, two questions can be asked. First, what is the capability of an electric propulsion system for a given range of values of these parameters, and how do such electric propulsion systems compare with the present and anticipated chemical propulsion systems? Second, for a given mission, what would be desirable values for these parameters?

MISSION DEFINITION

Satellites and probes, both manned and unmanned, can be used for a variety of purposes, *e.g.*, scientific investigation, meteorology, communication relays, and navigation, to mention a few. The first step for any of these is to boost the vehicle into space. This must be done chemically, because the thrust for electric propulsion systems is too low. The second step consists of one or more orbit changes. For some purposes, a third step, landing on Earth or some other body, may be required.

For missions considered in this paper, namely the Earth-Mars and Earth-Venus missions, it will be assumed that the vehicle starts from a circular orbit about the Earth at an altitude of 200 miles. The first class of missions will consist of a transfer from this close-in orbit about Earth to satellite orbits about Mars and Venus. The other class of missions will include the return from the target planet to a 200-mile orbit about Earth. A simplified model will be used. It will be assumed that all the orbits are coplanar and that perturbing forces will be neglected. The mission parameters considered are the duration of the mission and the payload fraction.

Because a particular mission can be accomplished many ways, the problem of optimization arises. Some of the possible objectives are to minimize the total radiation dosage, cost, or take-off weight per unit payload weight. For the above missions, the optimization criterion will be to maximize the payload fraction. The power source will not be considered a part of the payload.

The electric propulsion system considered will be power limited, and it will be assumed that the rate of mass flow, the specific impulse, and the direction of thrust can be regulated, subject only to the power limitation. The use of such an idealized propulsion system is of interest for the following reasons. First, no propulsion system can give a better payload fraction then the idealized system with the same maximum power. This gives an upper bound for the performance of electric propulsion systems. Second, a standard is set which shows how much is lost because of the use of a less flexible propulsion system. This gives some indication as to whether the development of more flexible engines should be considered.

The powered mission will be considered as consisting of two types of maneuvers. One of these will be the escape from or entry into a circular orbit about a planet, and the other will consist of a transfer from the orbit about the sun of the starting planet to the orbit about the sun of the planet which is the objective. The escape or entry maneuver will be regarded as occurring in an inertial frame, *i.e.*, a coordinate system, in which Newton's laws hold, and whose origin is the center of the planet considered. This maneuver will consist of imparting a velocity to the vehicle, starting from a circular orbit about a planet, which is sufficient to enable it to escape from the gravitational influence of that planet. At the termination of this maneuver, it will be assumed that the effect of the gravitational force of the planet is negligible, and that the velocity of the vehicle relative to the planet is zero. For the entry maneuver it will be assumed that the vehicle is at rest, relative to the planet, at a distance at which the gravitational force of the planet is negligible. The second maneuver will be regarded as occurring in an inertial frame whose origin is at the center of the sun. Each maneuver will be considered as occurring in the gravitational field of a single body which is a point mass. Since the mass of each body whose motion is of interest is negligible by comparison with the mass of the body whose field is considered, the field will be considered as a central field, whose center is at the attracting body, and for which the field force per unit mass or acceleration due to gravity is

$$a_r = \frac{-\mu}{r^2}, \tag{16}$$

$$a_\theta = 0, \tag{17}$$

where r and θ are polar coordinates with the origin at the center of the field, a_r and a_θ are the radial and circumferential components of the acceleration, and μ is the numerical value of the acceleration at unit distance from the center of the field. The orbits of Venus, Mars, and Earth were taken as being coplanar and circular with radii which would give the accepted values for their energies. The method of calculation utilized is readily applicable to a much more precise model, which includes perturbing forces, noncircular orbits, etc. However, the comparative capability of electric and chemical propulsion systems can be evaluated from this simplified model and will be adequate for the purposes of this paper.

The performance of chemical propulsion systems will also be calculated using the above model. In addition, it is assumed that the thrust force for the chemical system is impulsive, *i.e.*, that the time integral of the thrust per unit mass is the change in vehicle velocity. This means that the burn-out time is taken as being sufficiently small so that the change in position is negligible during this time. With this assumption, the velocity increment in the direction of thrust due to the propulsion system is given by (4). The velocity increment required will be such that the velocity of the vehicle, after the escape is effectively completed, is the velocity in the transfer orbit. The impulse at the point of intersection of the transfer orbit with the orbit of the target planet will be sufficient to change the orbit of the vehicle from the transfer orbit to the orbit of the target planet, and to place the vehicle in a circular orbit about the target planet. The velocity increment will be used as a measure of the mission requirements for chemical propulsion systems.

For electric propulsion systems, another parameter was suggested as a measure of mission requirements by Irving and Blum,[1,2] in connection with their elegant variational formulation of the thrust-programming problem. By (1) and (2)

$$M^2 f^2 = \dot{M}^2 C^2,$$

where f and C are the scalar magnitudes of $\bar{f}$ and $\bar{C}$. If this equation is divided by (8) the result may be written

$$-\frac{\dot{M}}{M^2} = \frac{f^2}{2P}.$$

Let M_0 and M_1 be the initial and terminal masses respectively. An integration yields

$$\frac{1}{M_1} - \frac{1}{M_0} = \int_0^T \frac{f^2}{2P}\,dt.$$

Since f and P may be varied independently, M_1, and thus the payload, will be a maximum if P is always as large as possible. For the power-limited engine, this implies that P is fixed at its largest value. Hence

$$\frac{M_1}{M_0} = \left[1 + \frac{M_0}{P} J\right]^{-1}, \tag{18}$$

where

$$J = \frac{1}{2}\int_0^T f^2 dt. \tag{19}$$

The value of J used for a mission is the value obtained for a thrust program which accomplishes the desired mission and maximizes M_1/M_0. In terms of the specific mass, (18) becomes

$$\frac{M_1}{M_0} = \left(1 + \frac{M_0}{M_w}\alpha J\right)^{-1}. \tag{20}$$

Since

$$M_1 = M_L + M_S + M_w, \tag{21}$$

it follows that the payload fraction

$$\frac{M_L}{M_0} = \frac{M_1}{M_0} - \frac{M_w}{M_0} - \frac{M_s}{M_0},$$

or by (20)

$$\frac{M_L}{M_0} = \left(1 + \frac{M_0}{M_w}\alpha J\right)^{-1} - \frac{M_w}{M_0} - \frac{M_s}{M_0}.$$

If the payload fraction is regarded as a function of (M_w/M_0), then a straightforward calculation shows that the payload fraction is a maximum when

$$\frac{M_w}{M_0} = \sqrt{\alpha J} - \alpha J,$$

and the maximum value is

$$\frac{M_L}{M_0} = (1 - \sqrt{\alpha J})^2 - \frac{M_s}{M_0}. \tag{22}$$

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[1] J. H. Irving, "Low-thrust flight: variable exhaust velocity in gravitational fields," in "Space Technology," H. Seifert, Ed., John Wiley and Sons, Inc., New York, N. Y., pp. 10-01 to 10-54; 1959.

[2] J. H. Irving and E. K. Blum, "Comparative Performance of Ballistic and Low-Thrust Vehicles for Flight to Mars," presented at the Second Annual AFOSR Astronautics Symp., Denver, Colo.; April, 1958.

This analysis is valid only for the idealized engine for which the specific mass is independent of the maximum power level. An analogous procedure has been developed for the case where the specific mass is a function of the power level.[3] The evaluation will be based on the assumption that (M_w/M_0) is optimum, thus permitting the use of (22).

To include the effect of the structural factor, (15) is solved for M_s/M_0:

$$\frac{M_s}{M_0} = K\left(1 - \frac{M_L}{M_0}\right), \tag{23}$$

and the result is substituted in (22). Solving the resulting equation for the payload fraction yields

$$\frac{M_L}{M_0} = \frac{(1 - \sqrt{\alpha J})^2 - K}{1 - K} \tag{24}$$

for the electric propulsion system.

For the chemical engine

$$M_1 = M_L + M_s, \tag{25}$$

and by (4)

$$\frac{M_L}{M_0} = \frac{M_1}{M_0} - \frac{M_s}{M_0} = e^{-\Delta v/C} - \frac{M_s}{M_0}.$$

By substituting in this equation the value of M_s/M_0, given by (23), the value of the payload fraction is found to be

$$\frac{M_L}{M_0} = \frac{e^{-\Delta v/C} - K}{1 - K} \tag{26}$$

for the chemical engine. The methods for calculating optimized values for Δv and J for various missions are given in the Appendix.

SUMMARY AND CONCLUSIONS

The first step in ascertaining the capabilities of chemical propulsion systems is to obtain the minimum velocity increment as a function of trip duration for the mission under consideration. Fig. 2 presents such minimized velocity increments for transfer from a satellite orbit around the Earth to a satellite orbit around the target planet. For all round-trip missions concerned, the unpowered waiting time in a satellite orbit at the target planet is not included in the mission durations shown. The details of the methods of calculation utilized in deriving these data are those shown in the Appendix.

In a similar manner, when electric propulsion systems are considered, minimized values of the acceleration integral J must be derived. Values of J derived for individual mission maneuvers are shown in Fig. 3.

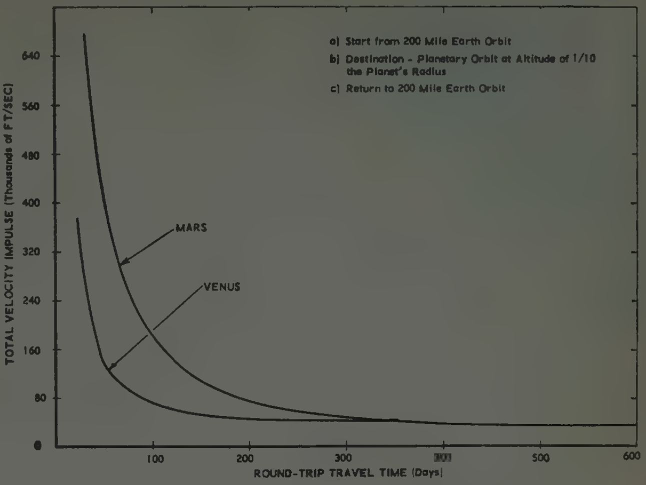

Fig. 2—Total velocity impulse required as a function of round-trip travel time for Mars and Venus missions. 1) Start from 200-mile Earth orbit; 2) Destination, planetary orbit at altitude of 1/10 the planet's radius; 3) return to 200-mile Earth orbit.

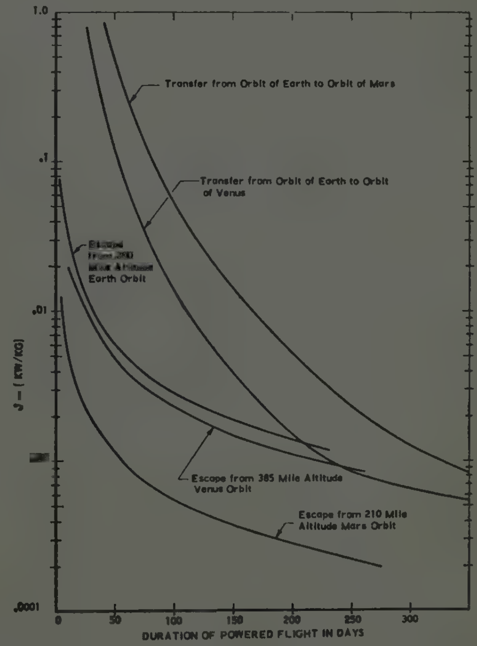

Fig. 3—Values of J required vs duration of powered flight for several maneuvers.

These maneuvers include escape from a 200-mile circular orbit about Earth, escape from circular satellite orbits about Venus and Mars, and transfer from the heliocentric orbit of Earth to the heliocentric orbits of Mars and Venus. The values of J and T, for entry into a satellite orbit, are the same as those for escape, and the mission parameters for transfer between heliocentric orbits are independent of the direction of transfer. Fig. 3 thus provides sufficient information about the various maneuvers which use low-thrust propulsion, so that the requirements for a complete mission may be ascertained. It should be noted that the values of J presented are optimum values for the heliocentric transfers and close to optimum for the planetary escapes and entries. It is also of interest to note that the escape maneuver requirements for J are of the same order of magnitude as the heliocentric transfer requirements.

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[3] "Advanced Space Propulsion Systems, Mission Analysis and Propulsion System Matching," New Devices Labs., Thompson Ramo Wooldridge Inc., Cleveland, Ohio, WADC Tech. Rept. 59-365; July, 1959.

From the results for the individual maneuver the optimized value of J for a complete mission can be obtained; such results are shown in Fig. 4 for escape from an Earth satellite orbit to entry into a satellite orbit about Mars and Venus. The curves drawn are for one-way missions; for an optimized round-trip, the values of J and T are twice those of a one-way trip.

In order to compare the capabilities of electric and chemical propulsion systems, the payload fraction (M_L/M_0) which can be obtained for a given duration T has been chosen as the figure of merit. For chemical systems the payload fraction depends primarily on the specific impulse. For purposes of this paper, specific impulses of 300, 400, 500 and 1000 seconds were selected. A specific impulse of 300 seconds is typical of present-day propulsion systems, while 500 seconds can be regarded as an upper limit. A specific impulse of 1000 seconds can be regarded as possible for thermonuclear propulsion systems which use hydrogen as a propellant. These high specific impulses were chosen to provide a "more-than-fair" consideration of conventional propulsion techniques. It should be pointed out, however, that even if such impulses are attained, doing so at the structural factor (*i.e.*, $K=0.05$) utilized herein is most unlikely.

In considering electric propulsion systems, the specific mass α was used as parameter, and values of α from 5 to 50 kg/kw were considered. The results obtained for several missions are shown in Figs. 5 through 8. It is obvious in these results that electric propulsion systems can provide outstanding mission capabilities for sufficiently low values of α.

A number of points are of interest in these payload fraction figures. First, an important difference is evident between electric and chemical systems. For chemical propulsion the payload fraction improves with increasing mission duration only up to a fixed point; beyond this point an increase in mission time decreases the payload. For electric propulsion, a longer mission always results in a larger payload, although the rate of increase is less for longer times.

Second, it is evident that for equal mission durations electric systems can meet or exceed the performance of 500-second chemical systems. This is even more pronounced for the longer round-trip missions. If somewhat longer durations for the same mission are allowed, the electric system far surpasses the chemical. For example, consider the Earth-Mars round trip in Fig. 7. Powerplant specific masses of 20 kg/kw are certainly feasible today, and such a powerplant roughly equals the payload capability of a 500-second chemical system at its optimum payload time. If the allowed mission duration is increased 50 per cent, the electric system can carry approximately three times more payload than the chemical system. If realistic chemical systems are considered (*i.e.*, 300- to 400-second impulse) they are not at all competitive, nor are chemical systems competitive if optimistic predictions on α are considered. (In the literature, power plant specific masses of 5 kg/kw are presented as feasible in the power range above 10 kw.)

A final point to be considered is the use of the electric power source for purposes other than propulsion, once the destination satellite orbit at a target planet has been established. If this power source is utilized for

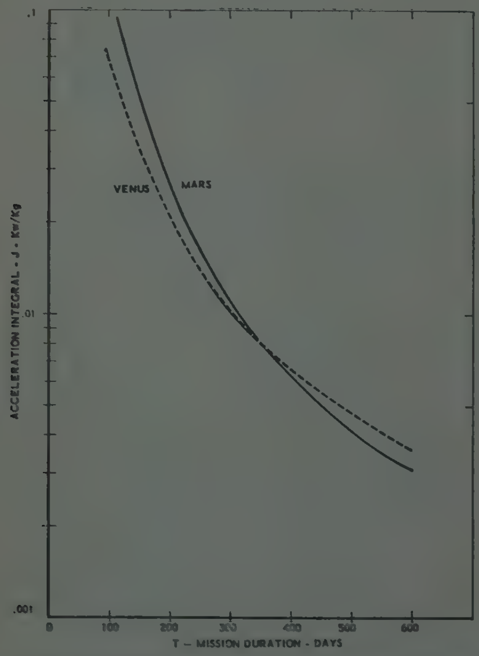

Fig. 4—Total acceleration integral J for an Earth satellite orbit-to-target planet satellite orbit mission, one-way trip.

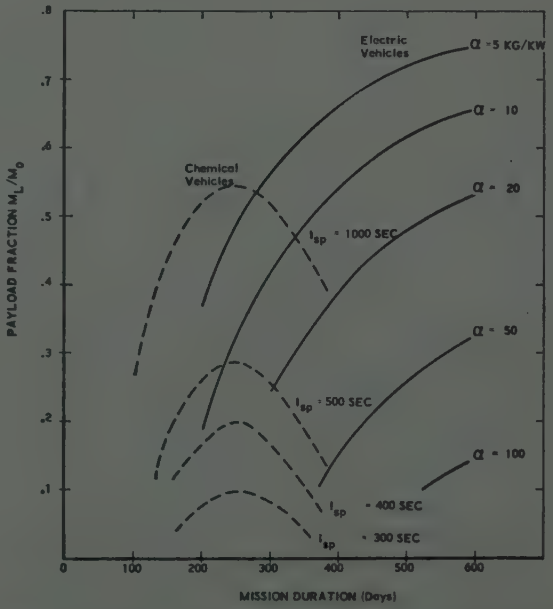

Fig. 5—Comparison of payload capabilities for electric and chemical vehicles on a one-way Mars mission. Structural factor, $K=0.05$.

communication, data processing, etc., it might be fair to include its weight, αP, in the payload fraction rather than in the propulsion system as in the previous figures. The results for such a calculation are given in Fig. 9. The payload fractions shown here compare to the ones previously shown in Fig. 5 for the identical mission. It is evident that such considerations emphasize the desirability of electric propulsion systems.

It should be noted that the acceleration integrals presented in Fig. 3 provide a means for the reader to construct payload capabilities for Venus and Mars missions other than the ones presented herein. For example, by entering a highly elliptical satellite orbit about Venus, an orbit whose eccentricity is nearly one, payload fractions much higher than those shown in Fig. 6 would be obtained. Entry into planetary satellite orbits other than those used in this paper can be determined by computing new J characteristics using the dimensional method of the Appendix.

APPENDIX

Orbits in an Inverse-Square Central Field

The results of this section are required for the calculation of the initial and terminal conditions, as well as the

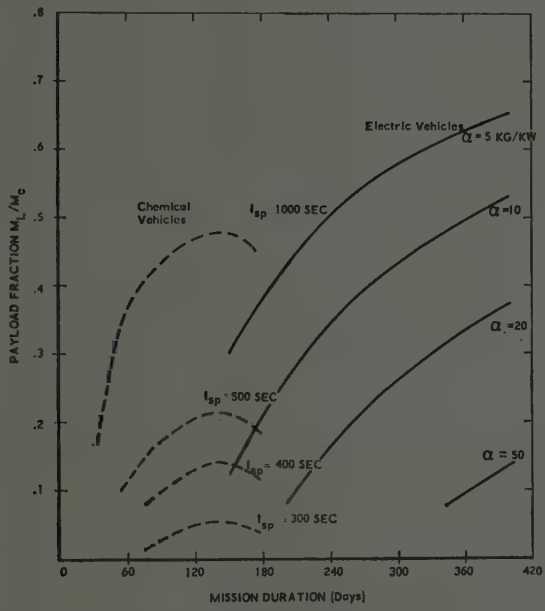

Fig. 6—Comparison of payload capabilities for electric and chemical vehicles on a one-way Venus mission. Structural factor, $K=0.05$.

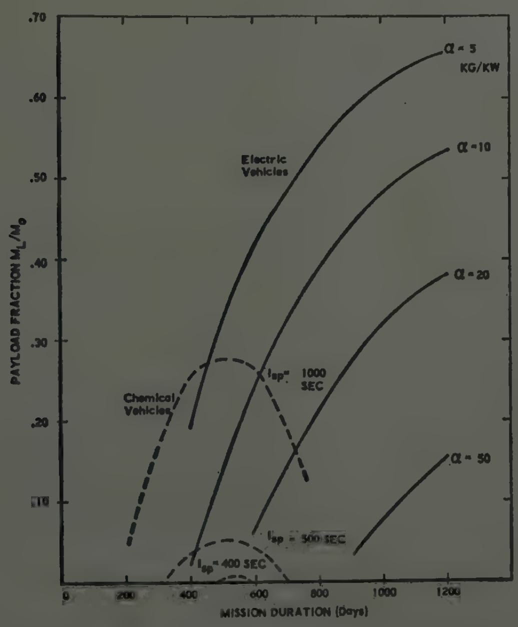

Fig. 7—Comparison of payload capabilities for electric and chemical vehicles on a round-trip Mars mission. Structural factor, $K=0.05$.

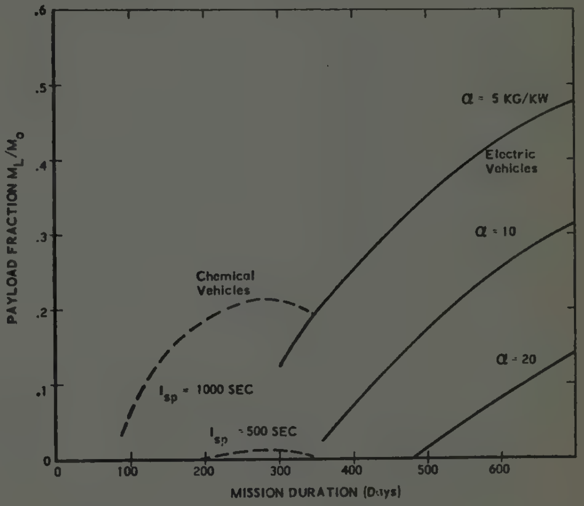

Fig. 8—Comparison of payload capabilities for electric and chemical vehicles on a round-trip Venus mission. Structural factor, $K=0.05$.

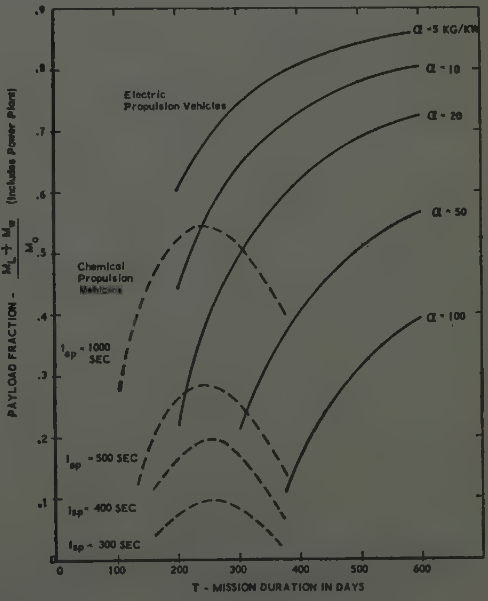

Fig. 9—Comparison of payload capabilities for electric and chemical propulsion vehicles on a one-way Earth-Mars mission where the electric power plant is considered part of the payload.

velocity increment and transfer time for impulsive orbit transfer.

In polar coordinates, the radial and circumferential components of acceleration are

$$a_r = \ddot{r} - r\dot{\theta}^2, \tag{27}$$

$$a_\theta = \frac{1}{r}\frac{d}{dt}(r^2\dot{\theta}). \tag{28}$$

By (16) and (17), the equations of motion for a particle whose mass is small with respect to the central mass are

$$\ddot{r} - r\dot{\theta}^2 = -\frac{\mu}{r^2}, \tag{29}$$

$$\frac{1}{r}\frac{d}{dt}(r^2\dot{\theta}) = 0. \tag{30}$$

Since the angular momentum of the particle per unit mass is defined by

$$h = r^2\theta, \tag{31}$$

(30) can be integrated immediately to yield

$$h = r^2\dot{\theta} = \text{constant}. \tag{32}$$

If (29) is written in the form

$$\ddot{r} - \frac{h^2}{r^3} + \frac{\mu}{r^2} = 0 \tag{33}$$

and multiplied by $\dot{r}$, it can be integrated directly to yield

$$\frac{1}{2}\left(\dot{r}^2 + \frac{h^2}{r^2}\right) - \frac{\mu}{r} = E, \tag{34}$$

where E is a constant of integration. Since the terms on the left are the kinetic and potential energy per unit mass respectively, E is the energy per unit mass, which is constant. Again, by (32)

$$\dot{r} = \frac{dr}{d\theta}\frac{d\theta}{dt} = \frac{h}{r^2}\frac{dr}{d\theta}.$$

Thus, (34) becomes

$$\left(\frac{dr}{d\theta}\right)^2 = \frac{r^4}{h^2}\left[2E + \frac{2\mu}{r} - \left(\frac{h}{r}\right)^2\right].$$

If the reciprocal of the radius is introduced as the dependent variable, this equation can be integrated by separation of variables to yield

$$r = \frac{h^2/\mu}{1 + e\cos(\theta - \theta_0)}, \tag{35}$$

where

$$e^2 = 1 + \frac{2Eh^2}{\mu^2} \tag{36}$$

and θ_0 is a constant of integration. This shows that the trajectory or orbit is a conic section with eccentricity e and focus at the center of the field. For $e=0$, the orbit is a circle; for $0<e<1$, the orbit is an ellipse; for $e=1$, the orbit is a parabola; and for $e>1$, the orbit is a hyperbola.

The velocity v in polar coordinates is given by

$$v^2 = \dot{r}^2 + (r\dot{\theta})^2, \tag{37}$$

and by (34) it is

$$v^2 = 2\left(E + \frac{\mu}{r}\right). \tag{38}$$

A particle is said to have escape velocity v_e if its trajectory is a parabola, *i.e.*, if $e=1$. By (36) this means that $E=0$. Hence, by (38) the escape velocity at a point is

$$v_e = \sqrt{\frac{2\mu}{r}}. \tag{39}$$

For a circular orbit, the circular velocity v_0 is obtained from (33) by noting that $\ddot{r}=0$. Thus the angular momentum h_0 in a circular orbit is

$$\frac{h_0^2}{r^3} = \frac{\mu}{r^2},$$

or

$$h_0 = \sqrt{\mu r}. \tag{40}$$

Since the velocity has no radial component,

$$v_0 = r\dot{\theta} = \frac{h_0}{r},$$

and hence

$$v_0 = \sqrt{\frac{\mu}{r}}. \tag{41}$$

It may be noted that (32) can be given another interpretation. The element of area swept out by a vector from the origin to the particle is

$$dA = \frac{1}{2}r^2 d\theta,$$

and hence

$$\frac{dA}{dt} = \frac{1}{2}h. \tag{42}$$

For elliptic orbits several parameters will be considered. The length a of the semimajor axis is half the sum of the minimum and maximum distances from the center of the field to the orbit. These distances are given when $\cos(\theta-\theta_0)$ assumes the values 1 and -1 respectively. Thus, by (35)

$$a = \frac{h^2}{2\mu}\left(\frac{1}{1+e} + \frac{1}{1-e}\right) = \frac{h^2}{\mu(1-e^2)}. \tag{43}$$

The length b of the semiminor axis is given by

$$b^2 = a^2(1 - e^2) = \frac{h^4}{\mu^2(1 - e^2)}\cdot \tag{44}$$

Since the area of an ellipse is πab, the period τ of a particle in an elliptic orbit is obtained by integrating (42). This gives

$$\tau = \frac{2A}{h} = \frac{2\pi ab}{h} = 2\pi\left(\frac{a^3}{\mu}\right)^{1/2}\cdot \tag{45}$$

The mean angular velocity m is

$$m = \frac{2\pi}{\tau} = \left(\frac{\mu}{a^3}\right)^{1/2}. \tag{46}$$

In cartesian coordinates with the x axis parallel to the major axis and the origin at the midpoint of the major axis, the parametric equations of the ellipse are

$$\begin{aligned} x &= a \cos \phi, \\ y &= b \sin \phi, \end{aligned} \tag{47}$$

where ϕ is the angle between the x axis and the vector from the origin of the cartesian coordinates to the particle. By considering the areas swept out by the radius vectors in both coordinate systems, the time t, required for the particle to move from a point closest to the center of the field to another point on its orbit, can be seen to be

$$mt = \phi - e \sin \phi. \tag{48}$$

The time of transit between any two points of the trajectory can be computed by using this formula two times. A similar formula using hyperbolic functions can be obtained for hyperbolic orbits. If (36) is solved for E, a comparison with (43) yields

$$E = -\frac{\mu}{2a}\cdot \tag{49}$$

This equation also holds for hyperbolic orbits provided a is allowed to assume negative values.

Impulsive Orbit-Transfer Between Circular Orbits

To change orbits impulsively at a point requires that the particle velocity $\bar{v}_0$ in the first orbit be changed impulsively to the velocity $\bar{v}_1$ required in the second orbit. Let E_0, h_0, and E_1, h_1, be the energies and angular momenta per unit mass in the first and second orbits respectively. Since no work is done by the field forces in an impulsive change of velocity, the work done which equals the change in kinetic energy is

$$\Delta E = E_1 - E_0 = \tfrac{1}{2}(v_1^2 - v_0^2) = \tfrac{1}{2}(\bar{v}_1 - \bar{v}_0)\cdot(\bar{v}_1 + \bar{v}_0). \tag{50}$$

If the velocity increment required is denoted by

$$\Delta\bar{v} = \bar{v}_1 - \bar{v}_0, \tag{51}$$

then

$$\Delta E = \tfrac{1}{2}\Delta\bar{v}\cdot(\bar{v}_0 + \bar{v}_1),$$

or

$$\Delta E = \tfrac{1}{2}(\Delta v)^2 + \bar{v}_0\cdot\Delta\bar{v}. \tag{52}$$

If the initial orbit is a circle, the angular momentum in the second orbit is the product of the radius and the component of $\bar{v}_1$ in the direction of $\bar{v}_0$, *i.e.*,

$$\Delta h = h_1 - h_0 = r\left(\bar{v}_1\cdot\frac{\bar{v}_0}{v_0} - \bar{v}_0\cdot\frac{\bar{v}_0}{v_0}\right),$$

or

$$\Delta h = r\left(\Delta\bar{v}\cdot\frac{\bar{v}_0}{v_0}\right);$$

hence

$$\Delta\bar{v}\cdot\bar{v}_0 = \frac{v_0\Delta h}{r}\cdot \tag{53}$$

Substituting for the last term in (52) yields

$$\Delta E = \frac{1}{2}(\Delta v)^2 + \frac{v_0\Delta h}{r}\cdot$$

This can be written

$$\left(\frac{\Delta v}{v_0}\right)^2 = \frac{2\Delta E}{v_0^2} - \frac{2\Delta h}{rv_0}\cdot \tag{54}$$

By (38) and (41) the energy per unit mass in a circular orbit is

$$E_0 = \frac{1}{2}v_0^2 - \frac{\mu}{r} = -\frac{1}{2}v_0^2,$$

and the angular momentum per unit mass in a circular orbit is

$$h_0 = rv_0.$$

Thus

$$\left(\frac{\Delta v}{v_0}\right)^2 = -\left(\frac{\Delta E}{E_0} + 2\frac{\Delta h}{h_0}\right). \tag{55}$$

This Δv is to be identified with the Δv in (4), under the assumption of negligible burn-out time. By changing the sign of the right side of (55), the velocity increment required to change from an arbitrary orbit to a circular orbit is obtained.

To illustrate the use of these formulas, the velocity increments needed for a Hohmann transfer orbit between the heliocentric orbit of Earth and the heliocentric orbit of Mars will be computed. The Hohmann transfer orbit is an ellipse which is tangent to the two circular orbits. Let r_e and r_m be the radii of Earth's and Mars' orbits, respectively, and let μ_s be the strength of the gravitational field of the Sun. The major semiaxis of the transfer orbit is

$$a = \tfrac{1}{2}(r_e + r_m). \tag{56}$$

By (49) the energy per unit mass of the transfer orbit is

$$E_h = -\frac{\mu}{r_e + r_m} \cdot$$

Since the velocity of the transfer orbit and the velocity of the initial orbit are parallel, $\bar{v}_0 \cdot \Delta\bar{v} = v_0 \Delta v$, and (52) becomes

$$\Delta E = \tfrac{1}{2}(\Delta v)^2 + v_0(\Delta v).$$

Hence

$$\frac{\Delta v}{v_0} = -1 + \sqrt{1 + \frac{\Delta E}{\frac{1}{2}v_0^2}} \cdot$$

But

$$\Delta E = -\frac{\mu_s}{2}\left(\frac{2}{r_e + r_m} - \frac{1}{r_e}\right)$$

$$v_0^2 = \frac{\mu_s}{r_e} \cdot$$

Thus,

$$\frac{\Delta v}{v_e} = -1 + \sqrt{\frac{2(r_m/r_e)}{1 + (r_m/r_e)}} \cdot \tag{57}$$

A similar calculation for the transfer to the Mars' orbit yields

$$\frac{\Delta v'}{v_m} = 1 - \sqrt{\frac{2}{1 + (r_m/r_e)}} \cdot \tag{58}$$

The escape from the satellite orbit about Earth and the transfer from the heliocentric Earth's orbit to the Hohmann orbit can be combined in the following way. It is assumed that the velocity imparted to the vehicle in the field of Earth is such that when the vehicle has effectively escaped, *i.e.*, when the potential energy of Earth's field is negligible, the remaining velocity is the velocity increment Δv required for the transfer to the Hohmann orbit. The energy of the vehicle, relative to Earth, will be $\frac{1}{2}(\Delta v)^2$, and the velocity v which the vehicle must have is, by (38),

$$v^2 = 2\left[\frac{1}{2}(\Delta v)^2 + \frac{\mu_e}{r_{oe}}\right],$$

where μ_e is strength of Earth's field, and r_{oe} is the radius of the circular satellite orbit about Earth. If v_{se} is the escape velocity from the orbit in Earth's field, then by (39) the required velocity is

$$v^2 = (\Delta v)^2 + v_{se}^2. \tag{59}$$

The vehicle has a velocity in the circular satellite orbit about Earth, which by (39) and (41), is $v_{se}/\sqrt{2}$. Hence, the total velocity increment Δv_e, required for escape from the close-in orbit and transfer to the Hohmann orbit, is

$$\Delta v_e = \sqrt{(\Delta v)^2 + v_{se}^2} - \frac{v_{se}}{\sqrt{2}} \cdot \tag{60}$$

For the transfer from the Hohmann orbit to the heliocentric Mars' orbit and entry into a circular satellite orbit about Mars, the velocity increment Δv_m required is

$$\Delta v_m = \sqrt{(\Delta v')^2 + v_{sm}^2} - \frac{v_{sm}}{\sqrt{2}} \cdot \tag{61}$$

The total velocity increment Δv_T required for the mission is:

$$\Delta v_T = \Delta v_e + \Delta v_m. \tag{62}$$

The time T required will be half of the period of the Hohmann orbit which is by (45) and (56)

$$T = \frac{\pi}{2}\left[\frac{(r_e + r_m)^3}{2\mu_s}\right]^{1/2} \cdot \tag{63}$$

For other transfer orbits (48) is used.

The value of Δv_T used for a given T is the smallest possible value. This was obtained by computing Δv_T as a function of T for several values of the transit angle. The envelope of this family of curves gave the minimum Δv_T for a given T. The time spent in satellite orbit about Mars and Venus is not included in these calculations. For round trips, Δv_T and T were doubled.

Escape and Entry Using Low Thrust

Studies made by Lawden,[4] Benney,[5] Tsien,[6] and Irving[1] indicate that constant circumferential thrust per unit mass is very close to optimum, and constant tangential thrust per unit mass is nearly as close to optimum. The figures used for escape from earth are those obtained by Irving and Blum.[2] For a mission starting from a close-in orbit about the earth, and ending in a close-in orbit about the target planet, the value of J is the sum of the values of J for the individual maneuvers. Since there are no dissipative terms, the reversal of the thrust program for escape gives the thrust program for entry. From the graph of J_e as a function of T_e, the thrust acceleration f can be computed from

$$J_e = \tfrac{1}{2}\int_0^{T_e} f^2 dt = \tfrac{1}{2}T_e f^2. \tag{64}$$

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[4] D. F. Lawden, "Optimal escape from a circular orbit," *Astronaut. Acta*, vol. 4, no. 3, pp. 218–233; 1958.

[5] D. J. Benney, "Escape from a circular orbit using tangential thrust," *Jet Propulsion*, vol. 28, pp. 167–169; March, 1958.

[6] H. S. Tsien, "Takeoff from satellite orbit," *ARS Journal*, vol. 23, pp. 233–236; July–August, 1953.

The equations of motion for constant tangential thrust per unit mass are

$$\ddot{r} - \frac{h^2}{r^3} + \frac{\mu}{r^2} = \frac{f\dot{r}}{v}, \tag{65}$$

$$\frac{1}{r}\frac{dh}{dt} = \frac{fh}{rv}, \tag{66}$$

where

$$v = \sqrt{\dot{r}^2 + \left(\frac{h}{r}\right)^2}. \tag{67}$$

The initial conditions are those for a circular orbit:

$$t = 0, \qquad r = r_0, \qquad \dot{r} = 0, \qquad h = \sqrt{\mu r_0}. \tag{68}$$

At $t = T_e$ the terminal condition is

$$E = E_1 = \tfrac{1}{2}\left[\dot{r}_1^2 + \left(\frac{h_1}{r_1}\right)^2\right] - \frac{\mu}{r_1} = 0, \tag{69}$$

where the subscript 1 refers to the values at escape. The solution of this problem is obtained by integrating the equations of motion numerically, and monitoring the value of the energy.

If the radius of the circular orbit is taken as unity and the period in the circular orbit is taken as the unit of time, the value of μ is, by (45),

$$\mu = 4\pi^2.$$

The initial conditions become at $t=0$

$$r_0 = 1, \qquad \dot{r}_0 = 0, \qquad h_0 = 2\pi,$$

and the terminal condition remains $E_1 = 0$. The equations of motion retain their form. In these units,

$$J_e = J_e(T_e) \tag{70}$$

is independent of the radius of the initial circular orbit, as well as the strength of the central field; hence, the solution for escape from Earth can be used to give the escape or entry from or into a circular orbit, using constant tangential thrust per unit mass, for any central field. A direct calculation yields

$$J_m = \sqrt{\left(\frac{g_{om}}{g_{oe}}\right)^3\left(\frac{r_{om}}{r_{oe}}\right)}\,J_e, \tag{71}$$

$$T_m = \sqrt{\left(\frac{r_{om}}{r_{oe}}\right)\left(\frac{g_{oe}}{g_{om}}\right)}\,T_e, \tag{72}$$

$$J_v = \sqrt{\left(\frac{g_{ov}}{g_{oe}}\right)^3\left(\frac{r_{ov}}{r_{oe}}\right)}\,J_e, \tag{73}$$

$$T_v = \sqrt{\left(\frac{r_{ov}}{r_{oe}}\right)\left(\frac{g_{oe}}{g_{ov}}\right)}\,T_e, \tag{74}$$

where T_e, T_m, and T_v are the escape times; J_e, J_m, and J_v are the values of one-half the integral of the square of the thrust per unit mass; g_{oe}, g_{om}, and g_{ov} are the accelerations due to gravity at the orbits around the planets; and r_{oe}, r_{om}, and r_{ov} are the radii of the initial orbits about Earth, Mars, and Venus, respectively. Since the altitude of the Earth orbit is 200 miles and the other orbits are at altitudes of one-tenth of the radii of the planets, the above formulas become

$$\left.\begin{aligned} J_m &= 0.154J_e \\ T_m &= 1.265T_e \\ J_v &= 0.708J_e \\ T_v &= 1.136T_e \end{aligned}\right\}. \tag{75}$$

The Approximation of J for Interplanetary Orbit Transfer

The calculation of J requires the determination of that trajectory

$$r = r(t), \tag{76}$$

$$h = h(t), \tag{77}$$

which makes

$$J = 1/2\int_0^T f^2 dt \tag{78}$$

a minimum, where

$$f^2 = f_r^2 + f_\theta^2, \tag{79}$$

$$f_r = \ddot{r} - \frac{h^2}{r^3} + \frac{\mu}{r^2}, \tag{80}$$

$$f_\theta = \frac{1}{r}\frac{dh}{dt}, \tag{81}$$

and, by (40),

$$\left.\begin{aligned} r(0) &= r_0 \\ \dot{r}(0) &= 0 \\ h(0) &= \sqrt{\mu r_0} \\ r(T) &= r_1 \\ \dot{r}(T) &= 0 \\ h(T) &= \sqrt{\mu r_1} \end{aligned}\right\}. \tag{82}$$

The conditions give the requirements for transfer in a central field from a circular orbit of radius r_0 to a circular orbit of radius r_1. Although the calculations have been formulated for the field due to a single body, this formulation can be extended to include the gravitational forces exerted by several bodies and other perturbing forces, as well as transfer between noncoplanar orbits. The transit angle can also be prescribed. In the form given above, the problem is a classical problem of the calculus of variations, and the Euler variational equations are a sixth-order system of nonlinear, ordinary differential equations. Some integrals have been obtained by Irving and Blum, but the solutions can be obtained only by numerical integration. This requires a trial and error process for determining three additional initial conditions, for which the particular solutions have the desired terminal values. In these calculations, the terminal conditions must be satisfied very closely, or large errors will result in the values of J obtained. Studies have indicated that J is extremely sensitive to terminal conditions.

The method used in this paper for the minimization of J is a direct method analogous to the Rayleigh-Ritz method. A family of trajectories which exactly satisfies the initial and terminal conditions is used, and J is minimized with respect to this family by a gradient method. Results obtained on a medium-size digital computer using this method have shown this procedure to be much more efficient and accurate than use of the Euler equations. The details of the method and the error analysis will be given elsewhere.[7]

The values of J and the transfer time selected for the total specified mission time T are those values for which the sum of the J's for the separate maneuvers is a minimum. This sum was obtained by first determining J_T+J_e as a function of T_e for various values of the parameter (T_e+T_T) where the subscripts e and T refer to Earth escape and heliocentric transfer from the Earth's orbit to Mars' orbit respectively. The minimum values of (J_T+J_e) for various fixed values of (T_e+T_T) were used in the same way with the Mars escape curve of J_m as a function of T_m to determine the minimum of $(J_e+J_T+J_m)$, for a fixed value of $(T_e+T_T+T_m)$. Similar calculations were made for the Venus trip. For round trips, the values of J and T were doubled.

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[7] C. Saltzer, "A variational method of determining optimum thrust program to accomplish coplanar orbit transfer," unpublished; 1959.

Electrostatic Propulsion*

STAFF OF THE RAMO-WOOLDRIDGE RESEARCH LABORATORY†

Summary—The thrust (newtons) developed by a system which accelerates a space charge limited flow of ions of specific charge q/m (coulombs per kilogram) through a voltage V is given by $F=(2\pi/q)\epsilon_0 V^2R^2=\sqrt{2m/q}\,V^2P$. R is an aspect ratio of ion beam diameter to acceleration distance x_0, and P is a diode perveance (amp/volts$^{3/2}$). Lines of electric force extend from charges in transit across the diode to charged surfaces of the accelerator electrode, on which the thrust is thus purely electrostatic. Alkali metals have favorable characteristics for use as propellants because their ions can be produced for small energy expenditure near 100 electron-volts per ion by surface contact ionization, while porous tungsten with pore size less than one micron is an efficient and convenient emitting surface. For practical values of thrust ($F>0.01$ lbf$=2.25\times10^{-3}$ newtons) and specific impulse ($I_{sp}>2000$ seconds) alkali metal thrust devices will operate at a few kilovolts and require $R\gg1$. But for $R\gtrsim3$, unipolar ion beams will reverse their direction of flow within a few x_0. Mathematical solutions to the neutralization problem for the case of one-dimensional flow of mixed ion and electron beams are presented assuming no energy exchanges by collisions or plasma instabilities. These show a periodic spatial potential distribution of wavelengths 0.027 x_0 for cesium, corresponding to time fluctuations of the potential at the plasma oscillation frequency in a frame of reference moving with the ions. Electrons having a Maxwellian distribution of velocity at 1000°K have an average velocity much faster than the ions, and for this case it appears that no time independent solution exists. The injection of electrons from strips spaced 2d apart and placed edgewise in the beam is analyzed under the hopeful assumption that there are no potential fluctuations along the direction of ionic motion. Electrons oscillate transversely across this direction in a potential trough having a depth midway between the cathode strips equal to $2V_0d^2/9x_0^2$. The electron current density is $75d/x_0$ times greater than the cesium ion current density. It is concluded that numerous closely-spaced electron injectors must be used, and that the total electron circulating current must be orders of magnitude greater than the ion current. Finally, if particles having much smaller specific charge than elemental ions could be copiously produced and accelerated, practically useful thrusts could be gained from pencil beam at $R\approx1$. The complications of neutralization would then probably be eliminated.

THE electrostatic thrust engine in its simplest form is a diode in which charged particles emitted by one electrode are accelerated and transmitted by another into space. If a voltage is applied in the absence of charged particle emission there is no current, no work is done, no power is required to maintain the potential difference, the electrostatic forces on the electrodes are equal and opposite, and there is no thrust. When we cause a current of ions to flow and be accelerated across a potential drop, power is expended and a thrust is exerted equal to the rate of change of momentum of the ions. The forces on the electrodes can no longer be equal and opposite; the detailed manner in which the unbalance occurs is shown by Fig. 1 and the equations which define space-charge-limited flow. For one-dimensional flow these are Poisson's equation,

$$\frac{d^2V}{dx^2} = -\frac{\rho}{\epsilon_0}, \tag{1}$$

the conservation of energy,

$$qV = \tfrac{1}{2}mv^2, \qquad \text{i.e., } v = \sqrt{2qV/m}, \tag{2}$$

and the equation of continuity,

$$J = nqv = \text{constant for all } x. \tag{3}$$

(MKS units will be used and the symbols are defined at the end of the paper.)

These lead to the familiar Child's Law,

$$J = \frac{4}{9}\epsilon_0\sqrt{\frac{2q}{m}}\,\frac{V^{3/2}}{x^2}. \tag{4}$$

The conservation of momentum defines the thrust per unit area,

$$\frac{F}{A} = \dot{M}v = Jvm/q \tag{5}$$

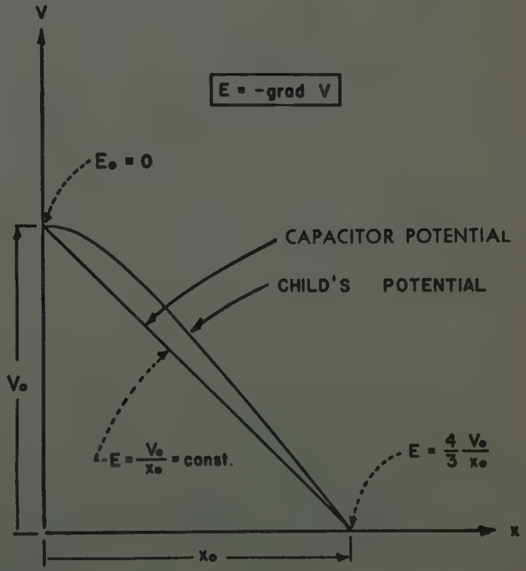

Fig. 1—The potential distribution in a parallel plane diode with and without space-charge-limited current flow.

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which can be combined with the preceding equation to yield,

$$\frac{F}{A} = \frac{8}{9}\,\epsilon_0\,\frac{V^2}{x^2}\,. \tag{6}$$

This last expression, except for a numerical factor of 16/9, is the electrostatic force per unit area on either plate of a parallel-plate capacitor,

$$\left(\frac{F}{A}\right)_{\text{capacitor}} = \frac{1}{2}\,\epsilon_0\,\frac{V^2}{x^2}\,.$$

Both the factor of 16/9 and the presence of an unbalanced force on one electrode are consequences of the nonuniform field along the direction of current flow. Fig. 1 compares the linear potential distribution across a capacitor with the curve distribution for space-charge-limited flow. The former has a constant gradient $E = -V_0/x_0$, but the field for Child's potential has a space dependence,

$$E(x) = -\frac{4}{3}\,\frac{V_0}{x_0}\left(\frac{x}{x_0}\right)^{1/3}.$$

The field is zero at the emitting electrode but is 4/3 stronger than a uniform field at the accelerating electrode,

$$E_{x=x_0} = -\frac{4}{3}\,\frac{V_0}{x_0}\,.$$

On squaring these fields, we find no force on the emitting electrode and the factor of 16/9 noted above in the force on the accelerating electrode. The thrust per unit area is simply an electric stress on an electrode caused by lines of force which terminate on the charged surface. At the other ends these lines terminate on space charge in transit. The thrust on an electric rocket is thus found by integrating the electrostatic pressures on the charged surfaces of its engine, in close analogy to the manner in which the thrust on a chemical rocket may be found from a complete inventory of pressure forces over the surfaces of its combustion chamber and nozzle.

ASPECT RATIO AND PERVEANCE

Eq. (6) can be put into the form,

$$F = \frac{8}{9}\,\epsilon_0 V^2\,\frac{A}{x_0^2} = \frac{2\pi}{9}\,\epsilon_0 V^2\left(\frac{D}{x}\right)^2 = \frac{2\pi}{9}\,\epsilon_0 V^2 R^2, \tag{7}$$

where $R = D/x_0$ is the "aspect ratio" of the diode. D is the diameter of a circle having the same area as the emitting region of the diode, and x_0 is the acceleration distance, *i.e.*, the diode spacing.

The total thrust of a space-charge-limited electrostatic system is therefore a function only of the total voltage and the aspect ratio of the beam. The relationships are shown in Fig. 2. While the total thrust is the same for all charge-to-mass ratios for a given V and R, the specific impulse (*i.e.*, the exit velocity)[1] is not. The latter must usually be set at a value which will meet requirements of optimum rocket performance. For an ion of given q/m, the voltage V is therefore fixed by considerations of the rocket mission and the power supply, as discussed elsewhere [1]. The value of the aspect ratio R is then determined by the required total thrust.

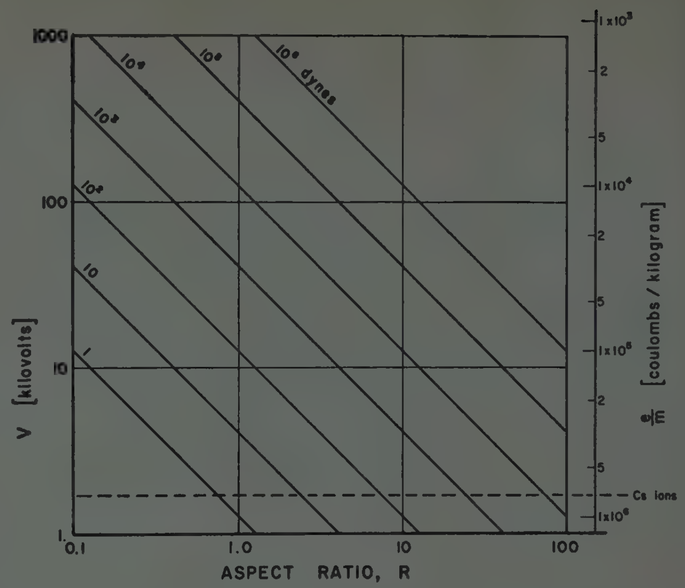

Fig. 2—Thrust as a function of the aspect ratio and the acceleration voltage. The number accompanying each line is the delivered thrust in dynes. The scale at the right indicates the correct acceleration voltage for a specific impulse of 5000 seconds for a given charge-to-mass ratio in the particles to be accelerated. A dashed line indicates that for Cs ions ($e/m = 7.5 \times 10^5$ coulombs per kilogram) an acceleration voltage of ~1.7 kilovolts is desired for a specific impulse of 5000 seconds.

The aspect ratio is closely related to the perveance, a familiar vacuum tube parameter defined as follows:

$$P \equiv I/V^{3/2}\ \text{amps}/(\text{volts})^{3/2}.$$

For a space-charge-limited diode,

$$P = \frac{\pi}{9}\,\epsilon_0\sqrt{\frac{2q}{m}}\,R^2 \begin{array}{l} = 1.83 \times 10^{-6} R^2 \text{ for electrons,} \\ = 0.37 \times 10^{-8} R^2 \text{ for Cs ions.} \end{array} \tag{8}$$

This equation relates the purely electrical quantity P to the geometrical parameter R. While an ionic thrust device will seldom, if ever, actually have a plane parallel configuration, the aspect ratio provides a basis for comparison of any system having a known perveance with an equivalent plane diode. R is, specifically, the ratio of diameter to spacing of a circular plane diode which would have the same perveance as the system under consideration. Implicit in this definition is the fact that for the same q/m all space-charge-limited diodes of a given shape have the same perveance regardless of their actual size.

As discussed in a later section, ions in a broad beam injected into free space will come to rest and reverse their motion within a distance approximately equal to the spacing of the equivalent diode. This spacing is thus effectively the "turn-around distance," and is a property of any monoenergetic ion beam in which motion and potential variation occur only along a single direction. The aspect ratio may thus be regarded as the ratio of beam diameter to turn-around distance, and is a useful parameter in discussions of charge neutralization. It should be noted that when a beam is expanded the diameter and the turn-around distance increase in proportion, so that R is invariant.

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[1] Specific impulse equals exhaust velocity divided by the acceleration of gravity.

It is apparent from (7) and Fig. 2 that $R \gg 1$ for almost any practical ionic thrust system using alkali ions. On the other hand, $R<1$ for most ion and electron beams encountered in practice. For example, a cathode ray beam with a current of 1 milliampere at 10,000 volts has an aspect ratio of about 1/40. For a thrust of 0.001 pound, using 1000 volt ions, $R=25$ (ion velocity $=37$ km/second for cesium, specific impulse $=3800$ seconds, current $=0.084$ ampere).

PRODUCTION OF IONS

The optimum exit velocity for an ionic propulsion system requires accelerating voltages ranging from a few kilovolts down to less than 1000 volts for some applications, provided singly charged elemental ions are used [1]. If high propulsive efficiencies are to be obtained, the power expended in forming an ion must be a small fraction of that required to accelerate it. An ion source with an energy consumption prior to acceleration of about 100 electron volts per ion, or less, is therefore desired. Values in this range have been experimentally obtained from the ionization of alkali metals on hot surfaces of high work function [2]. While encouraging efficiencies from gas discharge devices have also been reported [3], only surface ionization of the alkalis is discussed below.

The phenomena of surface ionization were studied in early classic experiments [4], [8] by placing a hot tungsten filament in a region filled with the vapor of an alkali metal. If the surface work function is higher than the ionization potential of the alkali atoms, the atoms lose an electron to the tungsten when they are adsorbed. After a period of residence on the surface, they evaporate as positive ions if the applied electric field permits. If the work function becomes lower than the ionization potential, the ionization process is much less effective and most of the alkali evaporates as neutral atoms. Since the effect of adsorbed alkali is to reduce the work function of tungsten, the ion-emitting state can be maintained only when the tungsten surface is slightly covered (less than about $\frac{1}{2}$ per cent) with adsorbed alkali. Below this degree of surface coverage, the rate of ion evaporation is almost exactly equal to the rate of arrival of atoms from the vapor. The rate of ion emission will rise in exact proportion to the arrival rate if the vapor pressure is increased, but the surface coverage on the tungsten will also rise. If pushed too far, this effect will lower the work function too much. For a given tungsten temperature, there is thus a maximum ion-emission current density. Attempts to increase the current beyond this point, or to reduce the temperature, cause the surface to change to the atom-emitting condition. We denote by T_c the critical temperature in degrees Kelvin at which the transition between the predominantly atom-emitting and ion-emitting conditions occurs. The tungsten temperatures required for ion current densities J amperes/cm² were found by early workers in this field [4] [8] to be as follows:

For

$$\begin{aligned} &\text{cesium,} && \log_{10} J = 8.99 - 14{,}350/T_c; \\ &\text{rubidium,} && \log_{10} J = 9.15 - 15{,}700/T_c; \\ &\text{potassium,} && \log_{10} J = 6.37 - 11{,}300/T_c. \end{aligned} \tag{9}$$

Numerical values based upon these equations are tabulated in Langmuir [5], [6].

The energy consumption consists almost entirely of losses caused by thermal radiation and this increases much more slowly than the ion current as T_c is raised. The efficiency of ionization therefore improves at higher current densities. In references cited above measurements up to 0.002 ampere/cm² for cesium and somewhat less for potassium and rubidium are reported. The ionization energy requirement for cesium at this current density is about 1000 ev/ion. While extrapolation of (9) indicated that the energy consumption might be as low as 45 ev/ion at 0.1 ampere/cm², no actual experimental data were available. It was of interest to know whether such high-current densities could, in fact, be drawn with high ion-to-atom ratios, and also whether the efficiencies would be as predicted by extrapolation.

Direct measurements at high-current densities have been reported by Shelton, Wuerker, and Sellen [2]. The tungsten was a filament 0.002 inch in diameter, and the collectors consisted of three coaxial cylinders 2 mm in diameter.

Fig. 3 shows how the current density of emitted ions varied as the filament temperature is raised through the critical temperature. Fig. 4 shows the current variation with voltage for potassium ions. The result is a typical diode characteristic. Relatively high voltages and small spacings were required to overcome the space-charge current of massive ions, and the saturation is limited by vapor pressure (*i.e.*, the rate of arrival), not by surface temperature. In Fig. 5 the saturation current density, that is, the density in the presence of an excess of voltage and with tungsten temperature higher than T_c, is plotted vs the equilibrium vapor temperature. One can see that cesium, rubidium, and potassium ion current densities above 100 ma/cm² were achieved, and that the results are in good agreement with those expected from the known vapor pressures of the alkalis.

These results confirm that the ion-emitting properties of the tungsten surface which had been observed at

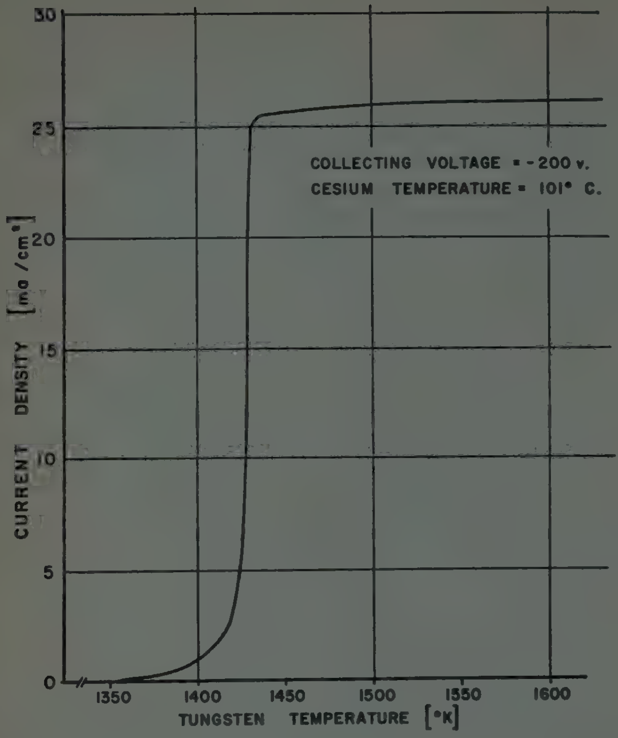


Fig. 3—The positive ion current emitted by a tungsten filament immersed in cesium vapor as the tungsten temperature is raised, showing the sharp transition from predominantly atom-emitting to ion-emitting condition at a critical temperature [2].

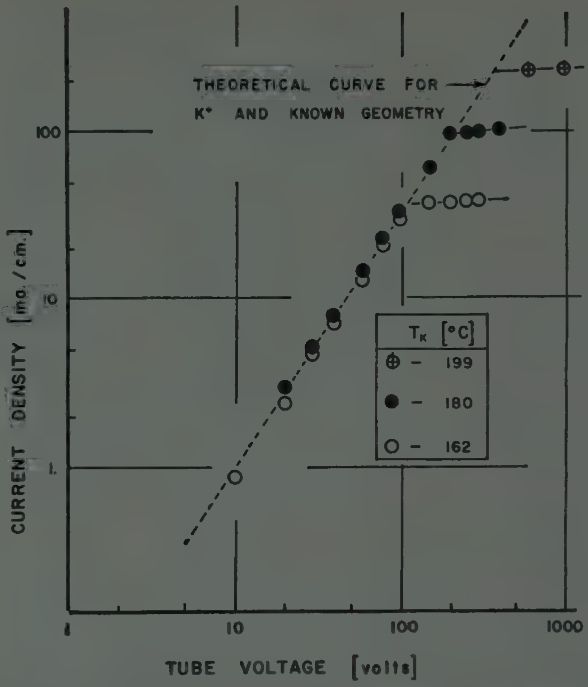


Fig. 4—Space-charge-limited behavior of potassium ions emitted by a tungsten filament immersed in potassium vapor. The tungsten was maintained at a temperature higher than the critical temperature T_c [2].

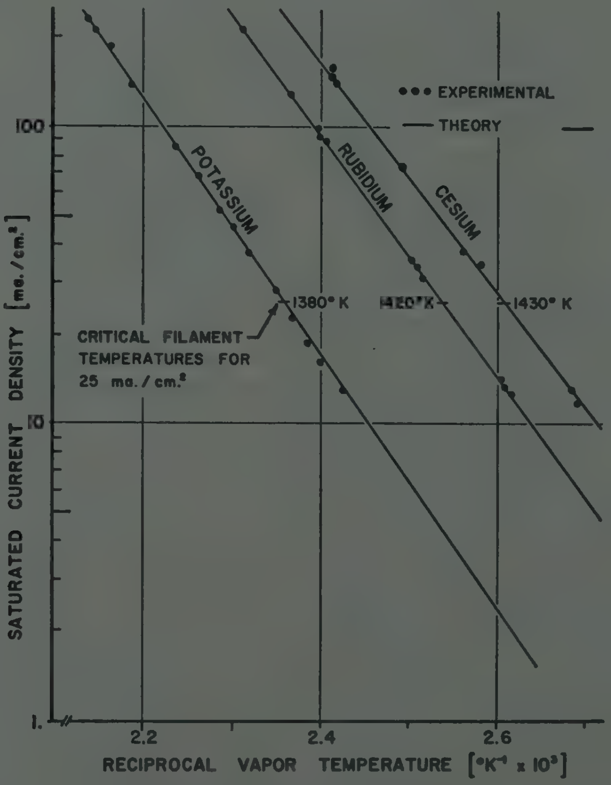


Fig. 5—Experimentally measured ion current densities from tungsten filaments [2].

low current densities function as expected for currents 1 to 200 times higher. Precise measurement of the critical temperature T_c at the higher ranges has not been possible, however, because of difficulties which plague experiments at high ion current densities. Temperature measurements on filaments as small as 0.002 inch are not satisfactory. Pyrometry is inaccurate because of filament vibration and measurements of temperature from the power input to the filament are uncertain because back bombardment by secondary electrons, which contributes significantly to the power, cannot be accurately estimated.

It is reasonably certain from the measurements that tungsten at 1450°K will support an ion current density of 25 ma/cm² with cesium and rubidium ions, giving an ionization energy consumption of 160 ev/ion, an acceptable value for practical purposes. The variation with temperature undoubtedly is close to that predicted from (9), so that values substantially below 100 ev/ion can doubtless be realized if the practical problems of handling higher current densities can be solved. While less than satisfactory from a scientific point of view, these statements are probably adequate for purposes of ionic propulsion in the present state of development. Practical thrust devices, as are discussed below, will use emitters much less simple than clean filaments in cylindrical geometry, and the value of T_c will be affected by changes in geometry, surface structure, and cleanliness. Moreover, problems more serious than exact knowledge of emitter temperature will dominate the field for a long time to come.

THE POROUS PLUG EMITTER

The line source in cylindrical geometry used in the experiments described above is not adapted for propulsion purposes. The production of an approximately unidirectional beam of ions with high mass flow rate raises serious problems in connection with supplying the neutral cesium to the hot surface. The electrical complications resulting from a high vapor pressure of cesium in the accelerating region are obvious. Also, leakage of neutral cesium from the exit apertures would cause appreciable wastage of mass. Such considerations have led many workers in the field to introduce the alkali vapor from the rear side of the emitter. Staggered ribbons, tungsten wire screens containing powders, and porous tungsten plugs have been tried [7].

The important questions are: 1) can an adequate mass flow be maintained at reasonable pressure; 2) can high enough current densities be drawn, and at temperatures which give ionization efficiencies comparable with those obtained using tungsten filaments; and 3) is the fraction of the alkali which escapes as neutral atoms sufficiently small?

Theoretical and experimental investigations of these questions will be described.

THEORY OF THE POROUS PLUG

We have cesium vapor proceeding from a supply at suitable pressure through a thickness of porous tungsten to an emitting surface, and wish to compute the fraction of atoms supplied which emerge as atoms, having failed to be ionized. We imagine the porous tungsten substructure to consist of long parallel tubes, of length L and radius a, with each tube servicing a plane emitting surface of radius $R=3a$. It will be convenient to study first the phenomena inside the tube, then the subsequent events on the emitting surface.

Inside the tube there are two phases of flow at the constant temperature of the tungsten, a volumetric phase of gaseous cesium atoms in the Knudsen regime and a surface phase diffusing along the wall under a concentration gradient. Wall coverage by the surface phase is never weak enough to permit appreciable internal ionization, which is therefore neglected. The two phases being coupled by the adsorption isotherm of Langmuir and Taylor [8], we can write Knudsen's equation for the flow rate I_v through the volumetric phase, the diffusion equation for the flow rate I_s through the surface phase, and form the quotient at $x=L$,

$$\left(\frac{I_v}{I_s}\right)_{x=L} = \frac{4}{3}\,\frac{a^2}{D_s\sigma_1}\exp\left\{62 - \frac{32380}{T}\right\}. \qquad (10)$$

This gives the proportion of emerging atoms which reach the end of the tube in the volumetric phase and are delivered un-ionized through the orifice. The balance continues to diffuse in the surface phase onto the disk-shaped emitting surface. In (10), σ_1 is the surface concentration of a complete monolayer of cesium on tungsten, $\sigma_1=3.6\times10^{14}$ atoms/cm². A word is required on the appropriate value of the temperature-dependent surface diffusion coefficient D_S, which is taken to be 2×10^{-4} cm²/second at 1600°K. A sequence of two delay periods determines this coefficient: the residence time at a site while awaiting escape from confinement in the adsorption potential, and the travel time to the adjoining site after escape is accomplished. While the trap-escape process depends exponentially on the temperature, the transit process depends more weakly on the square root of the temperature. At low temperatures the exponential dependence is controlling, but at about 1000°K the delay in transit comes to predominate. Accordingly, at $T=1600$°K, D_S is almost insensitive to temperature, and has a value smaller by an order of magnitude than would be found by extrapolating exponentially from the low-temperature data.

Before leaving the flow phenomena inside the tube, we should estimate the cesium vapor pressure which can supply a suitably strong surface phase to the ionization surface, and from that result in a current density of, for example, $J=0.01$ ampere/cm². By choosing a pore radius $a=0.2\ \mu$, we can make the proportion of neutral emission in (10) negligibly small, and by choosing $L=0.015$ inch and $T=1600°K$ to conform with practical experimental conditions, we can compute $p=5.0$ mm Hg from the flow equation. This value of cesium vapor pressure is in equilibrium with the liquid at 350°C, and is manageable in practice. It confirms the validity of application of Knudsen's equation in the argument leading to (10).

Finally, we turn back to the portion of the flow in surface diffusion at the end of the tube, which migrates onto the disk-shaped emitting plane. We have the data of Taylor and Langmuir for rates of atom and ion emission and their dependence on the fraction of surface covered, which fraction decreases from a maximum at the tube outlet to a minimum at the outer periphery of the disk. Because of the space-charge limitation on ion emission, the ion current density is uniform. The atomic emission is not electrically limited and varies radially with the coverage over the disk in a distribution which may be integrated to give the total atom emission from the surface of the disk. At a current density $J=0.01$ ampere /cm², these computations lead to Table I.

TABLE I

Pore Radius (Microns)		Percentage of Neutral Atoms Emitted	
		At 1500°K	At 1600°K
1.0	From Orifice	7	22
	From Disk	4–7	9–15
	Total	11–14	31–37
0.8	From Orifice	5	15
	From Disk	3–4	7–9
	Total	8–9	22–24
0.6	From Orifice	3	9
	From Disk	2–3	5–6
	Total	5–6	14–15
0.4	From Orifice	1	4
	From Disk	1–2	3
	Total	2–3	7
0.2	From Orifice	0.3	1
	From Disk	1	1
	Total	1	2

This compilation shows that the percentage of neutral atoms emitted falls rapidly as the pore radius is reduced from 1.0 μ, becoming acceptable at radii $a=0.2\ \mu$, and that it is advantageous to choose an operating temperature just above the critical temperature.

Experiment on the Porous Plug

Successful operation of porous tungsten emitters of cesium ions have been reported. Dulgeroff, Speiser, and Forrester [9] obtained a beam current of 0.0021 ampere, corresponding to a current density of about 0.012 ampere/cm² at the emitter, which was porous tungsten of 70 per cent theoretical density operated at 1575°K. Experiments by H. Shelton and J. M. Sellen have been directed toward measuring the fraction of neutral cesium atoms emitted from porous tungsten as the ion current density is increased. A heated tungsten probe is placed in the beam emitted from the plug and biased electrically so that it cannot receive the ionic component. Atoms incident on the probe are ionized, and the secondary ionic current from the probe is thus a measure of the atomic emission from the plug. At current densities of 0.0003 ampere/cm², the neutral emission from the plug was measured to be smaller than 5 per cent. Ionic current densities of 0.003 ampere/cm² have been drawn from a plug supplied at a cesium vapor pressure of 1 mm Hg, but, at this and higher densities, experimental circumstances prevented measurement of the associated atomic component. Special efforts were made to work at relatively low voltages of no more than a few hundred volts, but sputtering of cesium nevertheless occurred. Small interelectrode spacings were necessary to draw the higher current densities, and the full flow of ionic cesium collected and built up on the surfaces. Both the plug and the probe received substantial currents of atomic cesium sputtered from the electrodes.

It is believed that by choosing vapor temperature and plug temperature such as to keep the plug surface in the ion-emitting condition, much higher ion current densities can be drawn from porous tungsten plugs. Unambiguous proof of this and quantitative measurements of the atom-to-ion ratio will require great pains, particularly with respect to disposing of the mass transported by the ion beam and avoiding extraneous flows of cesium.

Charge Neutralization

The emission of many amperes of positive ion current from a space ship having a capacity of perhaps a millimicrofarad or less obviously imposes a severe requirement for maintaining the potential of the system in balance. A current of negative ions or electrons must be emitted in such a way as to meet simultaneously two separate conditions: 1) the net current from the space ship must be zero; 2) the space charge must be neutralized throughout the entire volume of the ion beam with sufficient accuracy to prevent serious loss of momentum or actual reversal of ion flow. The first condition presents no serious problem, since an electron beam can be ejected with the total desired current with small expenditure of energy and negligible transfer of momentum. But the neutralization of the ion beam throughout its volume appears to involve experimental and theoretical difficulties.

The following discussion first summarizes the behavior of unneutralized ion beams, and then presents theoretical results for some simple configurations of a neutralized beam for which analytical solutions can be found. While these idealized models do not adequately represent the real problem, they clarify some of the difficulties which must be overcome.

ONE-DIMENSIONAL THEORY OF UNNEUTRALIZED BEAMS [10]

Consider an emitter electrode at a positive potential V_0, an accelerating electrode at zero potential spaced at x_0, and a final electrode also at zero at a spacing x_1 from the accelerator. The current injected from the left is limited by Child's Law for the accelerating potential V_0 and the spacing x_0.

$$J = \frac{4}{9} \epsilon_0 \sqrt{\frac{2q}{m}} \frac{V_0^{3/2}}{x_0^2} .$$

In the space between the two electrodes at zero potential, the injected current produces a distribution of positive potential which depends on the ratio of the spacings x_1/x_0 in the injection and acceleration regions. As x_1 is increased from zero, the distributions are at first symmetrical about a maximum at the mid-distance $x_1/2$. The space is weakly charged and all of the incident current is transmitted. We will call these Class I potentials; they are sketched in Fig. 6, curves *a*, *b*, and *c*. As x_1 continues to increase and approaches 2.82 x_0, $V_{1,\max}$ approaches $\frac{3}{4}V_0$, a limiting case, shown by curve *c*, which is unstable against upward fluctuations of the incident current or of the spacing x_1. Should either of these occur, the distribution jumps discontinously and irreversibly to curve *d*, which is a Class II potential, associated with space-charge-limited current. The incident current is partially reflected and contributes twice to the local space charge in the region to the left of the potential maximum, at which a virtual source has formed. Class II distributions are then found as x_1 increases toward infinity and these finally result in total reflection of the incident current. The ions will come to rest and reverse their motion at a distance 0.707 x_0 from the accelerating plane. Let x_1 now be decreased. The space-charge-limited situation and the Class II potential distribution persists, the fraction of transmitted current increases, and the virtual source shifts to the right. At $x_1 = 2x_0$, we have curve *e*, a limiting case for which the fraction of ions reflected is zero. This condition is unstable against downward fluctutations of the incident current or the spacing x_1. Should either of these occur, total transmission is established and the potential jumps discontinuously and irreversibly to curve *f*, a Class I distribution for which $x_1 = 2x_0$ and $V_{1,\max} = \frac{1}{4}V_0$.

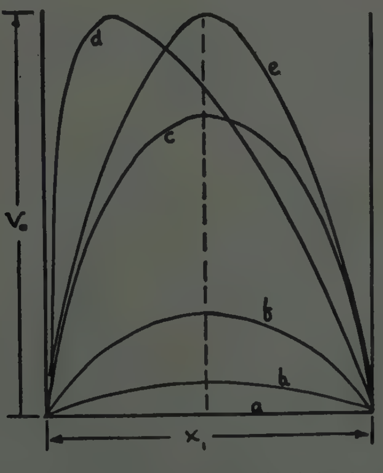


Fig. 6—Potential distributions produced by injection of monoenergetic ions into a region bounded by parallel planes at zero potential. The ions come to rest at potential V_0. The parameter which distinguishes the curves may be taken to be either the magnitude of the injected current density or the spacing x_0 to a third plane, to the left, which is held at potential V_0 and emits space-charge-limited ions.

Partial reflection cannot occur for $x_1 < 2x_0$, and total reflection must occur for $x_1 = \infty$, which is the case of an ion beam injected into infinite space. When $2x_0 < x_1 < 2.82x_0$, there is total transmission for Class I potentials and partial reflection for Class II potentials; but the two types of distribution cannot both exist together. Class I potentials are generated as x_1 increases through this interval; when the pattern of Class II is established, it persists while x_1 is returned through the interval in the decreasing sense. Practically, the largest value of x_1 which permits total transmission is just smaller than the instability at $x_1 = 2.82$ x_0, and must be reached in the increasing sense of x_1.

Neutralization must obviously be accomplished within a distance no greater then 2.82 x_0 from the injection plane. In a real configuration, as contrasted with the ideal plane diode, this requirement still applies if x_0 is taken as the "turn-around distance," the spacing of the equivalent diode which would produce the same current density at the same voltage as the ion beam under consideration. The actual beam might have been formed, for example, by the merging of a number of various beams, and a decelerating field might have been applied after the initial acceleration.

In sum, the foregoing analysis results in the following dilemma. In view of the dependence of thrust on aspect ratio, the requirement is for a broad beam with a large aspect ratio. But to broaden the beam indefinitely is to reduce it to one dimension, and a consequence of the one-dimensional analysis is the abrupt reflection of an unneutralized beam.

NARROW UNNEUTRALIZED BEAMS

The turn-around phenomenon described above in one-dimensional terms will occur only in a broad ion beam in which the diameter is considerably greater than the turn-around distance. If the aspect ratio $R \lesssim 1$, the beam will diverge but will not turn around [11]. The angle of divergence of a ray on the outside of the beam from the original direction of motion is given for small θ, by elementary considerations, as follows:

$$\theta \simeq 0.24R\left\{\ln\left[\frac{r}{r_0}\right]\right\}^{1/2},$$

where r is the radius of the expanded beam and r_0 is the initial beam radius. The contours of such beams are plotted in Fig. 7. The beams become essentially conical beyond a short distance ($\sim 10\ x_0$) from the orifice. Such beams will flow to very great distances from their source without turning back, even though no charge neutralization exists. Most beams of charged particles which are encountered in practical applications such as cathode ray tubes, mass spectrometers, and nuclear accelerators have aspect ratios much smaller than one. Beam power tubes such as klystrons and traveling-wave tubes require the highest perveance possible, and their beams are often prevented from expanding by longitudinal magnetic fields. Even in such cases, the maximum perveance achievable in a cylindrical beam corresponds to an aspect ratio of about 3 [11], the limit being established by the fact that ions on the axis of the beam are brought to rest by the space-charge repulsion. This represents the onset of the turn-around effect in the transition region of moderate values of R.

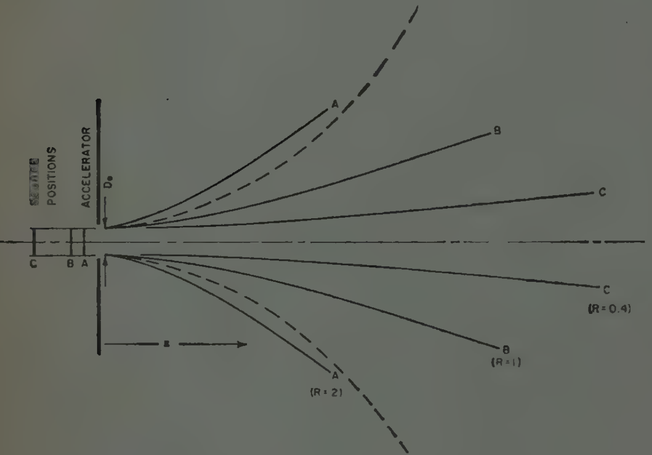


Fig. 7—Profiles of freely expanding beams for various aspect ratios (solid lines), and profile of an expanding slab beam (dashed line) whose thickness ($2y_0$) is equal to its acceleration distance (x_0).

SLAB BEAM

The behavior of a slab beam, narrow in the y direction but unlimited in z, can be calculated approximately from elementary considerations since the outward field E_y at all points along a given trajectory will be constant as long as the angle of divergence is small. If the width of the beam is $2y_0$ and the acceleration distance is x_0, the angle of divergence from the x axis of a ray at the outside of the beam will be

$$\theta \simeq 0.11\left(\frac{2y_0}{x_0}\right)\left(\frac{x}{x_0}\right).$$

The trajectory is a parabola. If the acceleration distance equals the beam width, the divergence will approach one radian at a distance of only about 10 acceleration distances, and the rate of increase will still be high. The behavior is completely different in magnitude from that of the pencil beam, and appears to offer little hope of relief from the essential problem of neutralizing high-perveance high-aspect ratio beams. With high-aspect ratios, it is essential that charge neutrality be achieved throughout the volume of the beam.

Recapitulating, we have discussed three possibilities for the unneutralized case: broad beams, pencil beams, and slab beams. The broad beams have large aspect ratios but turn around within a distance of the order of the equivalent diode spacing, as was shown from a one-dimensional argument. The pencil beams have small aspect ratios but a gentle conical divergence which enables them to flow to great distances without turning around. An intermediate class of slab beams offers moderate aspect ratios but has a bad divergence behavior.

The low aspect ratio beams can deliver appreciable thrust only when a large number are operated in parallel (*e.g.*, 2500 beams for each of which $R=1$ to give a thrust corresponding to $R=50$). But the suggested arrangement fails because of the interaction of the parallel beams. A group of parallel narrow beams are equivalent to a broad beam and will encounter the same turn-around difficulties.

THEORY OF NEUTRALIZED BEAMS IN ONE DIMENSION

We now take up neutralized beams, assuming that electrons are added to the ionic flow. The simplest case, in which the electrons have a unique initial velocity, is discussed first. The more realistic case of a thermal distribution of initial velocity is then examined. Finally a model in which electrons oscillate perpendicular to the direction of flow is discussed.

ELECTRONS EMITTED WITH UNIQUE VELOCITY—ONE-DIMENSIONAL FLOW [18]

If we define a potential V_e associated with the initial electron velocity $v_e(0)$, the conservation of energy yields

$$v_e(x) = \frac{2e}{m_e}\left[V_e + V(x)\right]^{1/2} = v_e(0)\left[1 + V(x)/V_e\right]^{1/2}.$$

We shall ignore variations in ion speed and take $v_p = (2eV_0/m_p)^{1/2}$. We also assume a current balance $J_p = \rho_p v_p = -J_e = -\rho_e v_e$, so that the vehicle will remain neutral. When these relations are substituted into Poisson's equation, we obtain

$$\frac{d^2V}{dx^2} = \frac{\rho_p}{\epsilon_0}\left[1/\gamma(x) - 1\right], \tag{11}$$

where

$$\gamma(x) \equiv v_e(x)/v_p = \gamma_0[1 + V(x)/V_e]^{1/2}.$$

If $\gamma_0 = 1$, corresponding to equal initial ion and electron velocities, we have ideal charge neutralization since $V = 0$ is the solution satisfying $V(0) = V'(0) = 0$. If $\gamma_0 \neq 1$, there will be variations in the potential which can be determined by integrating (11). The qualitative nature of these variations can be easily deduced. If the exit speed of the electrons is less than that of the ions, then $\gamma_0 < 1$ and $V''(0) > 0$. Since we require $V'(0) = 0$, the potential curve rises as x increases until the electron speed equals that of the ions at position x_i where $\gamma(x_i) = 1$ and the curve has an inflection. Since $V''(x)$ is positive for $0 < x < x_i$, this region contains an excess of negative charge. Beyond x_i, γ is greater than one, so $V''(x)$ is negative, and an excess of positive charge is present. The curve reaches its maximum at a position x_m such that the positive charge accumulated between x_i and x_m is just sufficient to cancel the negative charge between 0 and x_i. The potential at larger values of x symmetrically retraces itself until the electron speed reaches its original value $v_e(0)$ at $x = 2x_m$. We are thus led to a periodic solution of which the amplitudes at the inflection point and maximum are easily obtained from (11). Setting $V''(x)$ equal to zero yields

$$V_{\text{inf}} = (1/\gamma_0^2 - 1)V_e = \alpha^2(1 - \gamma_0^2)V_0$$
$$\alpha^2 \equiv m_e/m_p.$$

To calculate $V_{\max}$ we obtain the first integral of (11),

$$\left(\frac{dV}{dx}\right)^2 = \frac{2\rho_p}{\epsilon_0}\left\{\frac{2V_e}{\gamma_0}\left[(1 + V/V_e)^{1/2} - 1\right] - V\right\}, \quad (12)$$

and set V' equal to zero with the result,

$$V_{\max} = \frac{4}{\gamma_0^2}(1 - \gamma_0)V_e = 4\alpha^2(1 - \gamma_0)V_0,$$

so that

$$V_{\max}/V_{\text{inf}} = 4/(1 + \gamma_0).$$

For $\gamma_0 = 0$, *i.e.*, $v_e = 0$, $V_{\max}/V_0$ takes on its largest value $4\alpha^2 = 4m_e/m_p$ and $V_{\max}/V_{\text{inf}}$ has the value 4. As γ_0 approaches 1, $V_{\max}$ approaches zero and $V_{\max}/V_{\text{inf}}$ the value 2, the curve becoming a sinusoid of low amplitude.

A similar discussion can be given if the electrons are initially moving faster than the ions, up to a certain point. In this case, $\gamma_0 > 1$ and the potential distribution curves downward as x increases. A potential minimum at x_m has a depth given by the above formula for $V_{\max}$, which continues to be a valid result for the extremum as γ_0 increases from unity to larger values.

$$V_{\min}/V_e = -\frac{4}{\gamma_0^2}(\gamma_0 - 1).$$

This ratio becomes -1 when $\gamma_0 = 2$. The electrons are then brought to rest at x_m. For larger values of γ_0, a fraction r of the incident electrons is reflected and we must modify (11) which becomes

$$\frac{d^2V}{dx^2} = \frac{\rho_p}{\epsilon_0}\left[\left(\frac{1 + r}{1 - r}\right)\Big/\gamma(x) - 1\right], \quad (11a)$$

differing from the case of no reflection only by the presence of the factor $(1+r)/(1-r)$. Our previous formulas involving γ_0 and V_e *uncombined* will still be valid for the case of reflection if γ_0 is replaced by $\gamma_0(1-r)/(1+r)$. We now have

$$\left(\frac{1 - r}{1 + r}\right)\gamma_0 = 2$$

or

$$r = (\gamma_0 - 2)/(\gamma_0 + 2) \qquad \gamma_0 \geq 2.$$

Returning to the first integral of (11) and integrating again, the exact solution has been obtained by Nazarian for the case of no reflection, *i.e.*, $\gamma_0 \leq 2$,

$$x = \frac{2V_e^{1/2}}{\gamma_0(2\rho_p/\epsilon_0)^{1/2}}\left\{\cos^{-1}\frac{1 - \gamma_0(1 + V/V_e)^{1/2}}{1 - \gamma_0} \mp \frac{\gamma_0\left[\left(\frac{dV}{dx}\right)^2\right]^{1/2}}{(2\rho_p/\epsilon_0)^{1/2}V_e^{1/2}}\right\}, \qquad 0 < x < x_m,$$

with $(dV/dx)^2$ as given in (12). The upper sign is taken for $\gamma_0 < 1$ and the lower one for $\gamma_0 > 1$. We find the location of the inflection,

$$x_i/x_m = \frac{1}{2}\left\{1 - \frac{2}{\pi}(1 - \gamma_0)\right\}$$

and the spatial period λ,

$$\lambda = 2x_m = \frac{4\pi\alpha V_0^{1/2}}{(2\rho_p/\epsilon_0)^{1/2}}, \quad (13)$$

having used the relation $V_e^{1/2}/\gamma_0 = \alpha V_0^{1/2}$. Thus λ is independent of the initial electron velocity in the range $0 \leq \gamma_0 \leq 2$ where there is no reflection and completely periodic solutions exist. Two forms of expression for λ are of particular interest. First, using Child's law, we find

$$\lambda = 2^{1/2}3\pi\alpha x_0 = 13.3\alpha x_0$$
$$= 0.027x_0, \text{ for cesium.}$$

Thus, because of the factor α, ($\alpha_{C_s} = 2.02 \times 10^{-3}$), the space wavelength is much smaller than x_0 the distance through which the ions are accelerated. The second form of λ is obtained by expressing (13) in terms of v_p instead of V_0. Thus,

$$\lambda^2 = 4\pi^2 v_p^2/\omega^2_{\text{plasma}}, \qquad \omega^2_{\text{plasma}} = e\rho_p/\epsilon_0 m_e,$$

so that in a frame of reference moving with the ions there occur periodic variations with time of the potential with exactly the plasma frequency.

The exact solution for the case of reflection, *i.e.*, $\gamma_0 \geq 2$ is

$$x = \frac{V_e^{1/2}}{(2\rho_p/\epsilon_0)^{1/2}} \left\{ \cos^{-1} \left[2(1 + V/V_e)^{1/2} - 1 \right] + \frac{2\left[\left(\frac{dV}{dx}\right)^2\right]^{1/2}}{(2\rho_p/\epsilon_0)^{1/2} V_e^{1/2}} \right\}, \qquad 0 \leq x < x_m.$$

Comparing with the fixed value of x_m for the case of no reflection, we find

$$\frac{(x_m)_{\text{refl}}}{(x_m)_{\text{no refl}}} = \frac{1}{2} \gamma_0 \qquad \gamma_0 \geq 2.$$

At $(x_m)_{\text{refl}}$ the periodic solution corresponding to $\gamma_0 = 0$ joins on smoothly to the solution just found for the region $0 \leq x \leq x_m$, and starts off with the value $-V_e$. Fig. 8 shows a family of potential distributions generated as the parameter γ_0 varies from 0 to 2 which is the domain of no reflection, and beyond 2 where partial reflection must occur.

Electrons Emitted with a Maxwellian Distribution of Velocities [19]

With an arbitrary distribution $f(x, v)$ of electron speeds we have, for the forward current,

$$J_e = -e \int_0^\infty f(x, v) v dv = -e\bar{v}(x) \int_0^\infty f(x, v) dv = \bar{v}(x)\rho_e(x)$$

so that $\bar{v}(x)$ the mean speed in the distribution replaces the speed $v_e(x)$ of the unique velocity case. Thus, the condition for ideal neutralization is

$$v_p = \bar{v}(0).$$

For a Maxwellian distribution at the origin,

$$f(0, v_0) = A \exp(-m_e v_0^2/2kT) \qquad V_0 \geq 0,$$

the condition becomes

$$v_p = (2kT/\pi m_e)^{1/2}$$

If $\bar{v}(0) < v_p$, the argument for determining the steady-state potential proceeds exactly as for the unique velocity case except that we now speak of the *mean* electron speed. The velocity distribution is alternately shifted from lower to higher values of v, $\bar{v}(x)$ oscillating in value around v_p. The form of the distribution at various locations can be sketched as shown in Fig. 9(a). Periodic potential solutions have been obtained by an approximation method for this case of $\bar{v}(0) < v_p$ but will not be presented here.

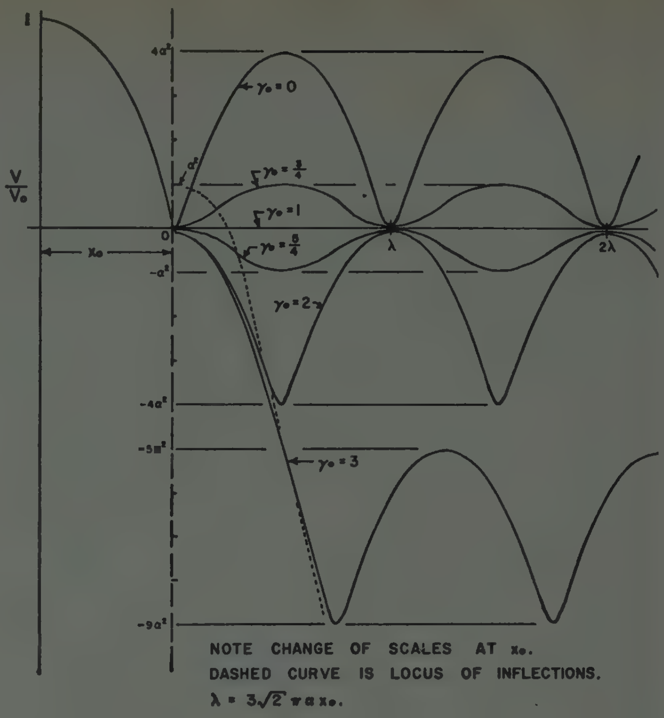
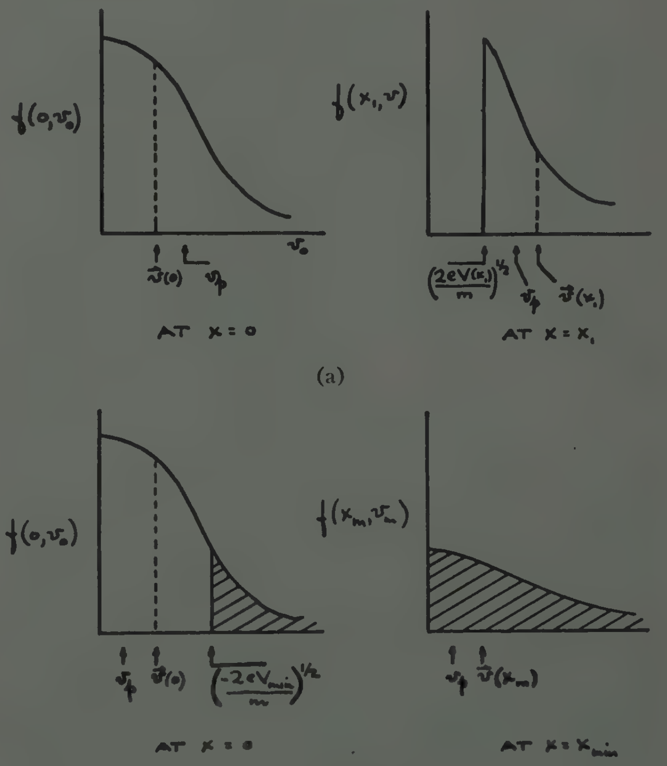


Fig. 8—One-dimensional potential distribution in an ion beam into which monoenergetic electrons have beeen injected parallel to the direction of ion flow. γ_0 is the ratio of initial electron to ionic velocity. Reflection of electrons occurs if $\gamma_0 > 2$. The wavelength $\lambda \ll x_0$.



(a)

(b)

Fig. 9—(a) Distribution function at different locations for the case $\bar{v}(0) < v_p$. (b) Distribution function at exit and potential minimum for the case $\bar{v}(0) > v_p$. The shaded portion is contributed by those electrons which eventually reach $x_{\min}$.

It might be thought at first that the situation for $v(0) > v_p$ is also similar to that of the corresponding unique velocity case, with the improvement that reflection can be interpreted properly as a separation of low-energy electrons from high-energy ones, rather than improperly by artificially attributing different behavior to electrons having identical energies. Arguing by analogy with the unique velocity case, we would expect the potential to fall as x increases and the electrons to be slowed down. However, this would result in continuous reflection of those electrons with sufficiently small initial velocities, and it is therefore the transmitted fraction of electron current which we must equate to the ion current leading to

$$\rho_e(x_m) = -J_p/v(x_m)$$

and we need $v(x_m) < v_p$ to give the potential a positive curvature. The question is whether v for the electron distribution at the minimum can be less than v_p as a consequence of the potential energy barrier, $-eV_m$ [Fig. 9(b)]. The answer to this question is in the negative since from Liouville's theorem we find

$$f(x_m, v_m) = A \exp(-m_e v_m^2/2kT) \exp(eV_m/kT) \quad v_m \geq 0$$

so that, in fact,

$$v(x_m) = v(0).$$

It is as though the potential barrier in the metal, corresponding to the electron work function, were increased and extended further out into space. The electrons would be emitted from this new "metal" in a Maxwellian distribution having precisely the same mean speed as the distribution emitted from the actual metal. The "work function" has increased by the amount $-eV_m$ with a corresponding decrease in the emitted electron current as expressed by the factor $\exp(eV_m/kT)$.

So if $v > v_p$ originally, then no matter how far the potential were to fall, v would remain unchanged. The only change in v/v_p would result from the accelerating effect of the field on the ions, which effect would be negligible in practice since $-eV_m$ must be of order kT or smaller to enable an appreciable fraction of the electron current to pass the barrier. We arrive at the important conclusion that if the electrons are emitted with a Maxwellian distribution of velocities, our one-dimensional model leads to stationary potentials only if at the exit plane the mean electron speed is less than the ion speed.

The condition will not be satisfied for most ion propulsion systems using alkali ions. At 1500°K, the mean electron speed will be 120 km/second, corresponding to a specific impulse of 12,000 seconds. A cesium ion voltage of 10 kilovolts would be required to match this electron velocity. These velocities are higher than the optimum values for many applications envisaged for ionic propulsion.

It is clear from the above discussion and from other obvious facts that the one-dimensional solutions of the neutralization problem have little if any relation to the practical case. The electrons cannot be introduced from an ideal plane source nor can their velocities be sufficiently small to satisfy the mathematical conditions postulated above. If the beam is not extremely well neutralized, positive potentials will develop which will accelerate electrons to speeds many times that of the ions. Multiple reflections of electrons will be essential to maintain current and charge equality.

A neutralization model which provides for multiple reflections of the electrons is that of transverse injection. Electrons are injected into the ion stream with initial directions which are perpendicular to the motion of the ions and oscillate back and forth across the ion beam under the influence of space charge fields. In order to provide equal ion and electron currents in the exhaust, the electrons can drift along the propulsive axis at a mean rate equal to the ion velocity. The behavior of electrons and ions in such a neutralization scheme leads to the simplified model of bipolar flow which is discussed in the following section.

NEUTRALIZATION WITH TRANSVERSE FLOW OF ELECTRONS

A second example of bipolar space and charge dynamics for which solutions may be easily obtained is the one-dimensional flow of electrons across an infinite slab of positive charge, as illustrated in Fig. 10(a). Electrons flow back and forth along the y direction from $y=0$ to $y=2d$, and the potential $V=0$ on the planes at these extreme positions. The positive charge is assumed to be fixed in space and of uniform density ρ_p in the region between $y=0$ and $y=2d$, and electrons are emitted with zero energy from these two planes. Poisson's equation, combined with others for continuity and conservation of energy, becomes

$$\frac{d^2V}{dy^2} = \frac{-\rho_+}{\epsilon_0} + \frac{J_e}{\epsilon_0\sqrt{\dfrac{2eV}{m}}},$$

where J_e is the total electron current density flowing in both directions; *i.e.*, the total current flux. If the total negative charge flowing back and forth is equal to the total positive charge in the slab, dV/dy vanishes at the boundaries, from Gauss' theorem. Solutions of the bipolar flow equation under these conditions are identical with those in (11) to (13) if γ is taken as the ratio of J_e (along y) to J_p at the point of maximum electron velocity at the bottom of the potential trough ($y=d$). The

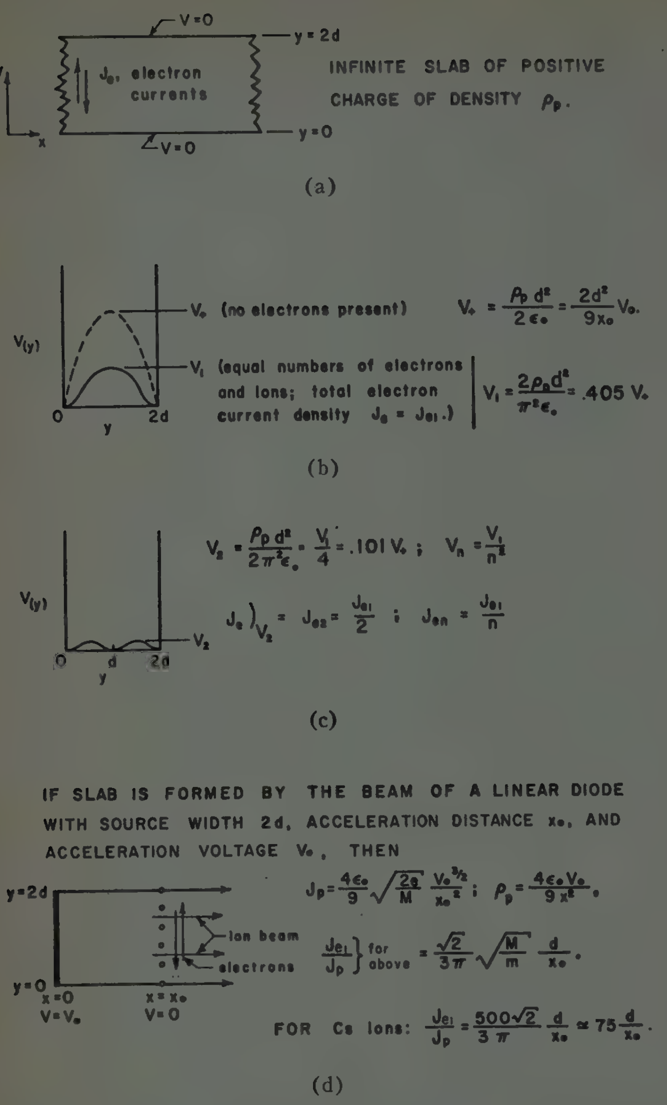

V=0
y=2d
J_e, electron currents
y=0
V=0
INFINITE SLAB OF POSITIVE CHARGE OF DENSITY ρ_p.
(a)
V(y)
0
2d
y
V_+ (no electrons present)
$V_+ = \frac{\rho_p d^2}{2\epsilon_0} = \frac{2d^2}{9x_0}V_0$.
V_1 (equal numbers of electrons and ions; total electron current density $J_e = J_{e1}$.)
$V_1 = \frac{2\rho_p d^2}{\pi^2\epsilon_0} = .405\ V_+$
(b)
V(y)
0
d
2d
y
V_2
$V_2 = \frac{\rho_p d^2}{2\pi^2\epsilon_0} = \frac{V_1}{4} = .101\ V_+$; $V_n = \frac{V_1}{n^2}$
$J_e\big)_{V_2} = J_{e2} = \frac{J_{e1}}{2}$; $J_{en} = \frac{J_{e1}}{n}$
(c)
IF SLAB IS FORMED BY THE BEAM OF A LINEAR DIODE WITH SOURCE WIDTH 2d, ACCELERATION DISTANCE x_0, AND ACCELERATION VOLTAGE V_0, THEN
y=2d
y=0
x=0
V=V_0
x=x_0
V=0
ion beam
electrons
$J_p = \frac{4\epsilon_0}{9}\sqrt{\frac{2e}{M}}\frac{V_0^{3/2}}{x_0^2}$; $\rho_p = \frac{4\epsilon_0 V_0}{9x_0^2}$.
$\left.\frac{J_{e1}}{J_p}\right\}$ for above $= \frac{\sqrt{2}}{3\pi}\sqrt{\frac{M}{m}}\frac{d}{x_0}$.
FOR Cs ions: $\frac{J_{e1}}{J_p} = \frac{500\sqrt{2}}{3\pi}\frac{d}{x_0} \approx 75\frac{d}{x_0}$.
(d)

Fig. 10—Summary of the properties of a slab beam of ions across which space-charge-limited electrons, oscillate transversely, assuming that no potential variations occur along the axis of the beam. To prevent some ions from coming to rest, the electron emitting planes must be less than about $3x_0$ apart, and J_e, the total transverse electron current density must be much larger than the ion current density, J_p.

ratio of electron to ion current densities is given by

$$J_e/J_p = \frac{\gamma_0}{2} = \frac{\sqrt{2}}{3\pi}\sqrt{\frac{M}{m}}\frac{d}{x_0} \tag{14}$$

and the depth of the potential trough is

$$V_1/V_0 = 0.090\,\frac{d^2}{x_0{}^2}\cdot \tag{15}$$

The properties of solutions of this type have been discussed by Naiditch, *et al.* [12].

A family of solutions results from the boundary conditions prescribed, two of which are illustrated in Fig. 10(b) and 10(c). The first consists of a single excursion in V, having maximum value V_1, as one moves from $y=0$ to $y=2d$. The second solution has only half the wavelength (*i.e.*, d), and an excursion in V which is $\frac{1}{4}$ of that for the first solution. The succeeding solutions have wavelengths of $2d/n$, and excursions in V of V_1/n^2. The parameter which is varied as one moves throughout this family of solutions is J_e, the electron current density. If the electron current density for which the first solution results is denoted as J_{e1}, the current densities for succeeding solutions with n wavelengths across the slab are given by J_{e1}/n. The required magnitude of J_{e1} relative to J_p, the ion current density, is indicated in Fig. 10(d).

The relationships presented in Fig. 10 permit preliminary estimates of the performance of a hypothetical transverse neutralizing system to be readily stated. Imagine a square uniform emitting area of side L, transmitting space-charge-limited ions from a source at acceleration distance $x_0 \ll L$. Let the electron emitters be ribbons mounted edgewise to the beam between which electrons travel transversely. Let the spacing between ribbons be $2d$, and the width of the ribbons be d. The total electron emitting area counting both sides of the ribbon is then L^2, for any value of d. The total electron current, for cesium ions, will be $75\ d/x_0$ times the ion current. By reducing d, the total electron current and the potential fluctuations in the neutralized beam will be reduced at the cost of added complexity of the ribbon structure. When $d=3.3x_0$, the analysis in Fig. 10 shows that $V_1=V_0$ and that the ions in the center of the beam would come to rest. Actually the assumption of uniform ion space charge makes the analysis not valid in this range, and the actual upper limit on d as set by onset of ion turn-around will be smaller than $3.3x_0$.

The treatment to this point has assumed that $V=V(y)$, and that no fields exist in directions other than the y direction. The placement of conducting planes at $y=0$ and $y=2d$ would provide that the conditions $\partial V/\partial x=0$ are, at least, met at the boundaries of the ion stream. The slab of charge, however, is neither infinite nor fixed, but is, in reality, a finite ion beam emerging from the acceleration region. With such an ion beam, axial electric fields exist which may be comparable with or larger than the fields in the y direction, and electrons, accelerated into the positive charge by these space-charge fields, may possess velocities along the axial direction of the same magnitude as their velocities transverse to the ion flow. The velocities of such electrons along the axial direction would be many times greater than ion velocities and the condition of electron drift velocities equal to the ion velocity would not be realized.

Thus, while the special situation of bipolar flow with spatial dependence in only the transverse direction provides revealing information about the qualitative nature of the neutralization problem, the existence of variations in V, ρ_+, and J_e in the x direction demands that the study of the motion encompass these complicating features if an adequate solution is to be found.

HEAVY PARTICLE PROPULSION

The problem of charge neutralization could be greatly simplified if particles of higher mass-to-charge ratio could be used. The difficulties of small electrode spacings would also be avoided, but a large number of new problems would also be introduced [20].

After converting the exit velocity to a specific impulse, we can combine (2) and (7) to give

$$F = \frac{\pi}{18} \epsilon_0 g^4 \left(\frac{m}{q} \right)^2 (I_{sp})^4 R^2.$$

For fixed aspect ratio and specific impulse, the thrust available from a beam is thus proportional to $(m/q)^2$. The advantage of high m/q lies in the opportunity to make R small. As pointed out in Fig. 7 and the associated discussion, a beam for which $R \sim 1$ will penetrate to a great distance into space, thus avoiding problems of turn-around of ions. If an equal electron current is projected in the same general direction, neutralization would presumably occur in a sufficiently remote, diffuse, and large region to make detailed mechanisms and space-charge disturbances irrelevant.

Fig. 2 shows the numerical values which apply if current is space-charge limited.

A method of charging and accelerating ion particles has been described by Shelton, Hendricks, and Wuerker [13]. The particles are charged by projecting them onto the surface of a conducting point maintained at a high potential. Theoretically, the maximum charge which can be placed on the particles is limited by the onset of field emission. If the limiting field strength is E_1, the minimum mass-to-charge ratio for a spherical particle of density ξ is

$$\frac{m}{q} = \frac{r\xi}{3\epsilon_0 E_1}.$$

The experimental value of E_1 is reported by Müller in connection with his work on the field emission microscope [14] at $E_1 \sim 10^{10}$ volts per meter for positively charged surfaces. The highest values of surface fields actually achieved and reported by Shelton, Hendricks, and Wuerker are about 2.5×10^9 volts per meter. This would correspond to a mass-to-charge ratio of 0.01 kilogram per coulomb on a 0.1-micron radius sphere, and to a velocity of 14 km/second ($I_{sp} = 1400$) with an accelerating potential of 10^6 volts. For higher I_{sp}, smaller particles would have to be used. In these experiments, the particle numbers were small, and present techniques are inadequate to produce a particle flux of sufficient magnitude to be useful.

A method which has been proposed for obtaining many particles with low m/q ratio is that of charging liquid droplets. Zelany [15] found that when the surfaces of liquids are subjected to high fields, they become unstable and eject sprays of small, charged droplets. At any small surface irregularity, the field strength is increased (fixed potential) and hence the tension of the surface due to the electric field is increased at the irregularity. The increase in tension further distorts the surface and the process continues until a point of liquid emerges from the surface. A fine spray of charged droplets issues from the liquid point. Schultz [16], Hendricks [17], and others have investigated charged liquid droplets as the possible working substance in an electrostatic, heavy-particle, thrust-producing device. Hendricks has measured diameters and charges of a sample of about 1000 droplets, finding many having a useful q/m, but a wide spread of lower values also.

In order to prevent the voltages from reaching unreasonably high values (*i.e.*, many megavolts) particles considerably smaller than one micron must be used.

A large extension of present techniques in electrical acceleration and ion optics will also be involved. While a value of aspect ratio R close to unity provides an attractively narrow beam compared to those required with elemental ions, it is very large relative to present practices in high voltage dc accelerators. A beam current of one milliampere at one mev has an aspect ratio of about 1/1000 for electrons, and 1/140 for protons.

CONCLUSION

Although electrical forces are sufficiently familiar, they have not yet been used to propel a space vehicle. The problems of efficiently producing and initially accelerating an ionic flow satisfactorily uncontaminated with neutral atoms appear to be essentially solved. Ion optics and sputtering, which have not been discussed in this paper, are problems which have not yet received adequate attention. Although their solution is vital to the achievement of adequate service life, they may be regarded presently as subordinate to the central problem of actually demonstrating significant thrust. The crucial importance of such a demonstration is pointed up by the following considerations: 1) all experiments thus far reported have been limited to low-aspect ratio, the perveances being two to three orders of magnitude smaller than those required for a practical device; 2) the experiments conducted at low perveance cannot be scaled up or operated in parallel because the high perveance beam raises problems different in kind from those encountered at low perveance; and 3) the problem of neutralizing a high perveance beam remains to be solved theoretically and experimentally.

NOMENCLATURE AND NOTATION

Italic Symbols

a = radius of a representative pore in a porous tungsten emitter;
A = area;
D = diameter, or representative length associated with the emitting area of an electrode;
D_s = the coefficient of surface diffusion for cesium on tungsten;
E, E_x, E_y, E_z = the electric field and its rectangular components;
F = thrust;
I = electric current;
g = acceleration of gravity;
I_v = volumetric portion of a gaseous flow rate;
I_s = adsorbed surface portion of a flow rate;
J = electric current density, amperes per unit area;
k = Boltzmann's constant;
L = length of a representative flow path through porous tungsten;
$\dot{M}$ = mass flow rate per unit area from an electrostatic thrust engine;
m, m_p, m_e = mass of a particle, and of an ion and electron, respectively;
n = numerical density of particles;
P = perveance;
q = charge per particle;
R = aspect ratio of a beam of particles, or transverse dimension divided by the acceleration distance;
r, r_0 = radius of a gently expanding pencil beam and its initial value; radius of a small sphere;
T = temperature, degrees Kelvin;
T_c = that temperature of an emitter at which the ionic emission predominates over the atomic emission;
v, v_p, v_e = velocity of a particle and of an ion and electron, respectively;
V = electric potential;
V_0 = the value of the potential drop across a diode accelerator;
V_e = the potential which is associated with an electronic velocity;
x, y, and z = Cartesian coordinates—x is taken as the direction of space dependence in all one-dimensional treatments;
x_0 = the spacing of a diode accelerator;
x_1 = a spacing downstream of a diode accelerator.

Greek Symbols

α^2 = ratio of electronic to ionic mass;
γ, γ_0 = ratio of electronic to ionic velocity, and its initial value, in the one-dimensional problem;
ϵ_0 = the permittivity of empty space;
λ = a spatial period or wavelength;
ρ, ρ_p, ρ_e = charge density, and density of ionic and electronic charge, respectively;
σ_1 = surface concentration of a monolayer of cesium on tungsten;
ξ = density of a small particle;
ω_{plasma} = the angular frequency of oscillations in a plasma.

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Photon Propelled Space Vehicles*

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Summary—**The interplanetary trajectories of vehicles propelled by solar radiation pressure are analyzed, and are shown to be logarithmic spirals if thrust direction is constant with respect to the vehicle-sun line. The required thrust may be obtained with a solar sail.**

Sail size as a function of trip time to Mars is determined for solar thrust, oriented tangent to the trajectory.

Solar propulsion is compared with chemical and electrical propulsion. It is shown that a solar-sail-powered space vehicle on a journey from earth to Mars operates with a payload and flight time penalty when compared with a ballistic vehicle. However, the work capacity per unit weight of a solar sail is calculated to be superior to an electrical engine, which in turn is vastly superior to a chemical engine when the work is compared on the basis of equal flight times.

INTRODUCTION

FORBES[1] showed that a logarithmic spiral is the space trajectory which results when thrust is applied tangent to the flight path, and when the magnitude of thrust is proportional to the inverse square of distance to the sun. Since the required thrust is small, the use of a solar sail is suggested. In this paper it is shown that the application of an inverse square thrust tangent to the flight path results in a special case of a more general class of logarithmic spiral trajectories.

Although some questions have been raised concerning the practical feasibility of solar sail vehicles for space travel, comparison with other continuous-thrust engines indicates that solar sailing must be given serious consideration in the light of present-day technology.

NOTATION

r = distance from sun to vehicle,
θ = longitude angle measured from the point where logarithmic spiral trajectory intersects the orbit of earth,
p = solar sail tilt parameter $=(\tan\phi)^{-1}=F_r/F_\theta$,
A = area of sail,
F_r = force per unit mass along radius vector,
F_θ = force per unit mass perpendicular to radius vector,
β = angle between tangent to spiral and radius vector,
q = constant in logarithmic spiral curve, $r=e^{q\theta}$, $=(\tan\beta)^{-1}$,
μ = gravitational constant of the sun,
S = solar constant,
m = total mass of vehicle,
t = time,
v = velocity.

Dots denote a differentiation with respect to time; subscripts o denote initial conditions; and subscripts f denote final conditions.

EQUATIONS OF MOTION

If the applied thrust is zero, and forces are resolved along a radius vector to the sun and perpendicular to the radius vector (circumferential), then the equations of motion in a central force field are:

$$\ddot{r}-r\dot{\theta}^2+\frac{\mu}{r^2}=0, \tag{1}$$

$$\frac{1}{r}\frac{d}{dt}(r^2\dot{\theta})=0. \tag{2}$$

When forces per unit mass F_r and F_θ are applied in the radial and circumferential directions respectively, then (1) and (2) become:

$$\ddot{r}-r\dot{\theta}^2+\frac{\mu}{r^2}=F_r, \tag{3}$$

$$\frac{1}{r}\frac{d}{dt}(r^2\dot{\theta})=F_\theta. \tag{4}$$

Assume that the trajectory is a logarithmic spiral of the form

$$r=e^{q\theta}, \tag{5}$$

where $1/q=\tan\beta=$ a constant. The geometry of the situation is described in Fig. 1.

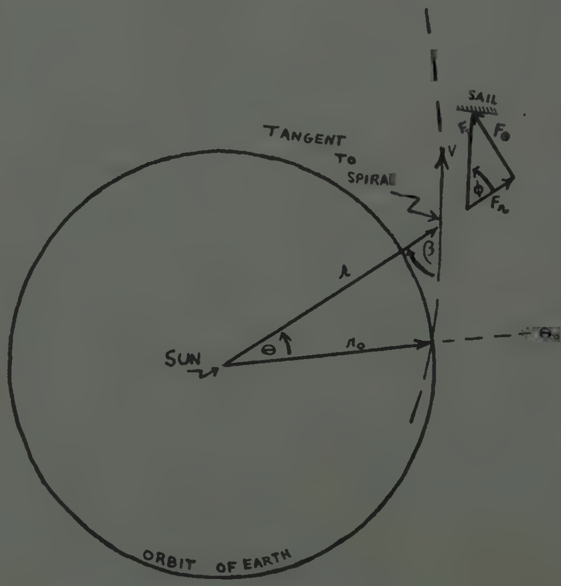

Fig. 1—Geometry of logarithmic spiral trajectory.

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Eqs. (3) and (4) may then be solved for r and θ as functions of time. Time differentiation of (5) provides:

$$\dot{r} = qr\dot{\theta}, \tag{6}$$

$$\ddot{r} = q\dot{r}\dot{\theta} + qr\ddot{\theta} = q(\dot{r}\dot{\theta} + r\ddot{\theta}). \tag{7}$$

After performance of the indicated differentiation in (4), F_θ is shown to be

$$F_\theta = 2\dot{r}\dot{\theta} + r\ddot{\theta} = (\dot{r}\dot{\theta} + r\ddot{\theta}) + \dot{r}\dot{\theta}. \tag{8}$$

Substitution of (8) into (7) yields

$$\ddot{r} = q(F_\theta - \dot{r}\dot{\theta}). \tag{9}$$

Now, $\ddot{r}$ in (9) may be set equal to the $\ddot{r}$ in (3) so that:

$$F_r + r\dot{\theta}^2 - \frac{\mu}{r^2} = qF_\theta - q\dot{r}\dot{\theta}. \tag{10}$$

Now let the sail apply forces per unit mass F_r and F_θ in a constant ratio p, such that

$$\frac{F_\theta}{F_r} = \frac{1}{p} = \tan \phi, \tag{11}$$

where ϕ is the angle between the sail normal and the radius vector to the sun. ϕ is also the angle of incidence of sunlight on the sail, as shown in Fig. 1. By combining (6), (10), and (11), we obtain:

$$F_\theta(p - q) + r\dot{\theta}^2(1 + q^2) = \frac{\mu}{r^2}. \tag{12}$$

Since solar pressure falls off as $1/r^2$, let $F_\theta = K_1/r^2$. Then (12) may be solved for $\dot{\theta}^2$, which yields

$$\dot{\theta}^2 = \frac{\mu + K_1(q - p)}{r^3(1 + q^2)} = \frac{\mu + K_1(q - p)}{1 + q^2} e^{-3q\theta}; \tag{13}$$

and, since

$$2\ddot{\theta} = \left[\frac{\mu + K_1(q - p)}{1 + q^2}\right](-3q)e^{-3q\theta}, \tag{14}$$

then

$$\ddot{\theta} = -3/2q\dot{\theta}^2. \tag{15}$$

From (8):

$$F_\theta = 2qr\dot{\theta}^2 - 3/2rq\dot{\theta}^2 = \frac{qr\dot{\theta}^2}{2} = \frac{K_1}{r^2} \tag{16}$$

or

$$\frac{K_1}{r^2} = \frac{q}{2r^2}\left[\frac{\mu + K_1(q - p)}{1 + q^2}\right], \tag{17}$$

and

$$K_1 = \frac{q\mu}{q^2 + qp + 2}. \tag{18}$$

Thus,

$$F_\theta = \frac{q\mu}{r^2(q^2 + qp + 2)}, \tag{19}$$

and

$$F_r = \frac{pq\mu}{r^2(q^2 + qp + 2)}. \tag{20}$$

Eq. (13) may be integrated to yield

$$e^{3/2q\theta} = 3/2q\sqrt{\frac{\mu + K_1(q - p)}{(1 + q)^2}}\,t + e^{3/2q\theta_o}, \tag{21}$$

where $\theta = \theta_o$ when $t = 0$.

Thus (5) and (21) yield

$$r^{3/2} = \frac{3q}{2}\sqrt{\frac{\mu + K_1(q - p)}{1 + q^2}}\,t + r_o^{3/2}. \tag{22}$$

TANGENTIAL THRUST

In the case where the force is applied tangent to the rocket path, then q equals p, $\beta = \phi$, and (19), (20), and (21), (22) respectively reduce to:

$$F_r = \frac{\mu p^2}{2r^2(p^2 + 1)}, \tag{23}$$

$$F_\theta = \frac{\mu p}{2r^2(p^2 + 1)}, \tag{24}$$

$$e^{3/2p\theta} = 3/2p\sqrt{\frac{\mu}{p^2 + 1}}\,t + e^{3/2p\theta_0}, \tag{25}$$

$$r^{3/2} = 3/2p\sqrt{\frac{\mu}{p^2 + 1}}\,t + r_o^{3/2}. \tag{26}$$

SAIL AREA AS A FUNCTION OF TRIP TIME

The case of thrusting tangent to the trajectory results in larger sail areas than necessary, but serves as a simple illustration of the method of determining the sail size. Later, we will show sail areas obtained for optimized tilt angles.

Fig. 2 represents the geometrical and force relationships of the sail with respect to the incident and reflected radiation from the sun.

A sail of area A is oriented so that light incident upon it makes an angle ϕ with the normal to the sail. The projected area of the sail, perpendicular to the incident radiation, is $A \cos \phi$. If S is the solar constant at Earth, and $S \cos \phi$ the radiation pressure normal to the sail, then the component of force in the radial direction must equal mF_r. Thus,

$$SA \cos^3 \phi = \frac{\mu}{2r^2}\,\frac{p^2\ m}{p^2 + 1}, \tag{27}$$

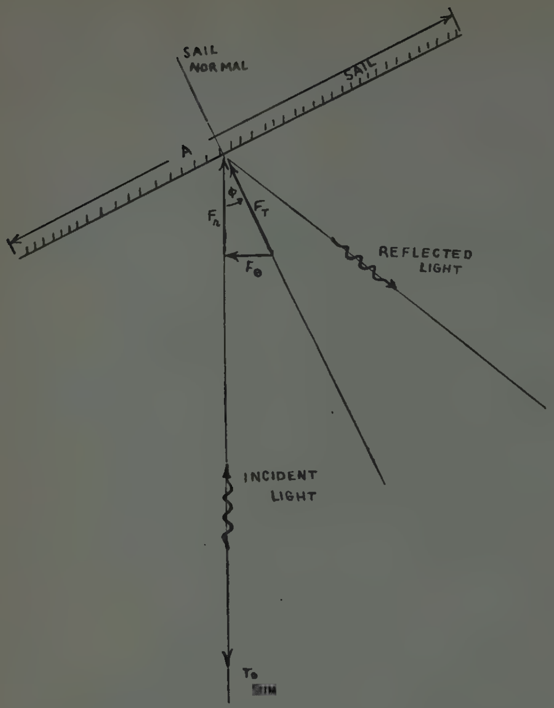

Fig. 2—Geometrical relationship of the sail to the incident and reflected radiation from the sun and resultant forces.

where r is now understood to be one astronomical unit and m is the total vehicle mass. Since $\tan \phi = 1/p$, then

$$\cos \phi = \frac{p}{\sqrt{p^2 + 1}}, \tag{28}$$

and

$$SA = \frac{\mu m}{2r^2}\left[\frac{(p^2 + 1)^{1/2}}{p}\right]. \tag{29}$$

The area of the sail required per unit total vehicle mass is

$$\frac{A}{m} = \frac{1}{S} \frac{\mu}{2r^2} \frac{(p^2 + 1)^{1/2}}{p}. \tag{30}$$

Since[2] S is 0.89×10^{-4} dynes/cm² at r equal to 1.50×10^{13} cm, and μ is 1.32×10^{26} cm³/sec², then

$$\frac{A}{m} = 0.33 \times 10^4 \frac{(p^2 + 1)^{1/2}}{p} \text{ cm}^2/\text{gm}. \tag{31}$$

In order to evaluate it as a function of trip time to Mars, rewrite (26) as

$$\frac{r_f^{3/2} - r_o^{3/2}}{3/2\mu^{1/2}} = \left[\frac{p}{(p + 1)^{1/2}}\right] t. \tag{32}$$

If the journey starts at Earth and ends at Mars, then

$$r_f = 1.524 \text{ AU}, \qquad r_0 = 1.000 \text{ AU}$$

and

$$\mu = 0.0002959 \text{ (AU)}^3/\text{day}^2,$$

and

$$\frac{(p^2 + 1)^{1/2}}{p} = \frac{t}{34.16 \text{ days}}. \tag{33}$$

The resultant area of sail required per unit mass is given in Table I in cm²/gram and in ft²/lb for various transfer times to Mars. The transfer angle, θ, measured from the start of the logarithmic spiral trajectory, is also listed.

TABLE I

AREA OF SAIL PER UNIT MASS (CM²/GM) OR PER UNIT WEIGHT (FT²/LB) AND TRANSFER ANGLE AS FUNCTIONS OF TRANSFER TIME FOR THRUST APPLIED TANGENT TO TRAJECTORY

Transfer time (days)	$\frac{p}{(p^2+1)^{1/2}}$	Sail area per unit mass or earth weight		Transfer angle, θ in radians
		$\frac{\text{cm}^2}{\text{gm}}$	ft²/lb	
80	0.427	77×10^2	37×10^2	0.9
100	0.342	97×10^2	47×10^2	1.2
150	0.228	140×10^2	70×10^2	1.8
200	0.171	190×10^2	94×10^2	2.4
250	0.137	240×10^2	110×10^2	3.1
300	0.114	300×10^2	150×10^2	3.7

The increase in sail area as journey time increases is explained in the term $\cos^3 \phi$ in (27). Although the required solar force decreases in magnitude as p decreases (*i.e.*, as the spiral angle ϕ increases), the corresponding reduction in effective sail area causes a net increase in required sail size. However, the energy necessary to inject the space vehicle into the spiral orbit is greater for shorter transfer times. Hence, optimization on a complete vehicle systems basis is indicated.

VEHICLE ACCELERATION

The total acceleration per unit mass (F_T) provided by the sail for the cases of tangential thrust may be determined from:

$$F_T^2 = F_r^2 + F_\theta^2 \tag{34}$$

and

$$F_T = \frac{\mu}{2r^2} \frac{p}{(p^2 + 1)^{1/2}} = 296 \frac{\text{cm}}{\text{sec}^2} \frac{p}{\sqrt{p^2 + 1}}. \tag{35}$$

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[2] C. D. Hodgman, Ed., "Handbook of Chemistry and Physics," U. S. Chemical Rubber Publishing Co., p. 2626; 1947.

This acceleration decreases as the vehicle travels farther from the sun. Table II lists the g's acceleration experienced by the sail in the vicinity of Earth and in the vicinity of Mars, as a function of transfer time.

TABLE II

ACCELERATION ON SAIL AS A FUNCTION OF TRANSFER TIME WHEN THRUST IS APPLIED TANGENT TO TRAJECTORY

Transfer Time (days)	g's at Earth	g's at Mars
80	13×10^{-5}	5.6×10^{-5}
100	10×10^{-5}	4.5×10^{-5}
150	6.9×10^{-5}	3.0×10^{-5}
200	5.2×10^{-5}	2.2×10^{-5}
250	4.1×10^{-5}	1.8×10^{-5}
300	3.4×10^{-5}	1.5×10^{-5}

THE APPLICATION OF NON-TANGENTIAL THRUST

From the foregoing equations, a sail tilt parameter, p, and hence the sail tilt angle, ϕ, may be computed, thus minimizing the sail area for a given trajectory (*i.e.*, for a given value of q). In general, these minima occur for tilt angles in the neighborhood of 30° to 40° with the optimum tilt angle increasing as β approaches 90°. Fig. 3 shows the sail area per unit weight plotted against the tilt angle ϕ for trajectories having β angles of 80° and 85°. It is immediately evident that sails sized for tangential thrust applications are excessive when compared to minimum sail areas, although the transfer times are slightly less. Fig. 3 also shows that the transfer time has a strong dependence upon the angle β and a weak dependence upon the angle ϕ. Thus, it is better to consider a trade-off between sail size and trajectory path than between sail size and sail tilt angle.

Fig. 4 shows that some control over flight time may be achieved by changing the tilt of the sail. By fixing a sail design point slightly above the minimum, the transit time may be increased or decreased by five to ten days with a small change in sail tilt. Of course, if the sail is originally set at minimum flight time for the given area and trajectory, then the transit time can only be increased by a sail tilt angle-change in either direction.

SOLAR SAIL PROPULSION COMPARED TO CHEMICAL PROPULSION

Because the photon-powered vehicle has a very low acceleration capability (typically 10^{-4} to 10^{-5} g's), it cannot leave the Earth under its own power. Some additional energy must be added to inject the vehicle into a logarithmic spiral trajectory. The speed of a vehicle in a logarithmic spiral trajectory is nearly the speed of a vehicle in a circular orbit at the same distance from the sun, but with a change in direction. After the vehicle has escaped from the gravitational field of the Earth, an additional increment of velocity, ΔV (see Fig. 5), must be applied to turn the velocity vector into the logarithmic spiral velocity requirement, where

$$\Delta V^2 = V_E^2\left[\left(1+\frac{2q^2+2}{q^2+qp+2}\right) - 2\sin\beta\left(\frac{2q^2+2}{q^2+qp+2}\right)^{1/2}\right]. \qquad (36)$$

If ΔV is applied after escape from the Earth, the total velocity budget needed to enter a logarithmic spiral is 6.97 mps (Earth escape velocity) plus ΔV. After escape

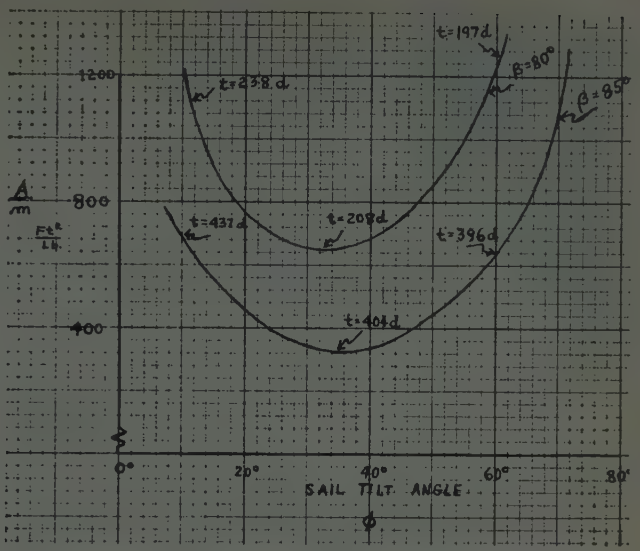

Fig. 3—Sail area required vs sail-tilt angle for trajectories from Earth to Mars.

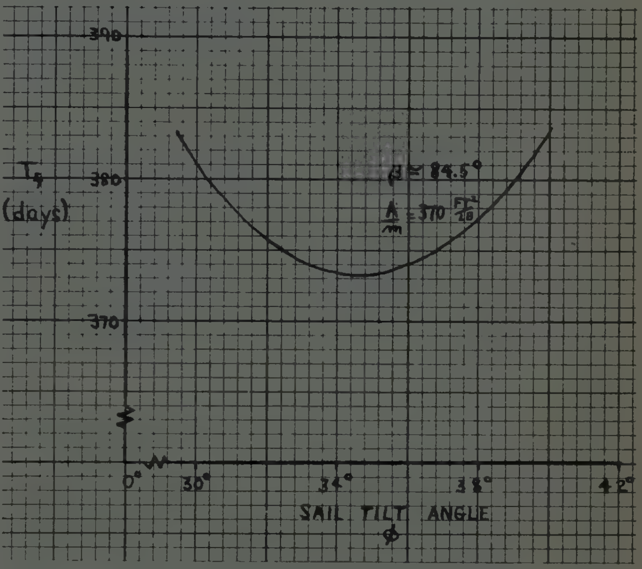

Fig. 4—Transfer-time vs sail-tilt angle, at constant sail area for trips from Earth to Mars.

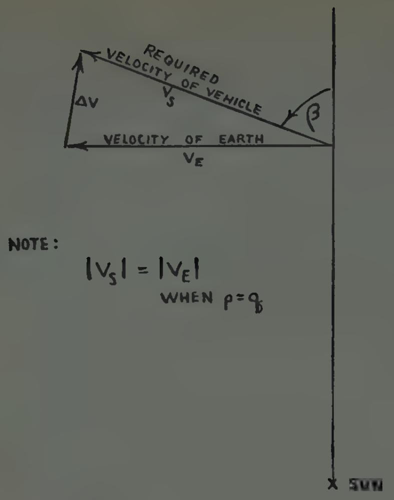

Fig. 5—Velocity coupling from orbital velocity to Earth to velocity of vehicle in log spiral.

the vehicle already has earth orbital velocity, and ΔV adjusts the direction. Fortunately, the burnout velocity, V_b, need not be the sum of 6.97 mps plus ΔV, if fuel is burned near the surface of the Earth. The minimum required burnout velocity (V_b) may be calculated by the following equation:

$$V_b^2 = 6.97^2 + \Delta V^2. \tag{37}$$

Then

$$\Delta V_n = V_b - 6.97, \tag{38}$$

where ΔV_n is the velocity increment needed, in addition to escape velocity, at burnout. Both ΔV and ΔV_b are listed in Table III.

TABLE III

Comparison of Solar Sail Vehicle with a Ballistic Vehicle for Flights to Mars

Injection angle β	ΔV (mps)	V_b (mps)	Injection payload (pounds)	Sail weight (pounds)	Transfer Time (days)	
					Solar sail vehicle	Ballistic vehicle
75°	4.8	8.5	580	110	143	94
80°	3.2	7.7	860	110	208	125
85°	1.6	7.2	1100	76	404	not possible

The same ΔV_n may be used to inject a ballistic vehicle of the same weight into an interplanetary (elliptical) trajectory. As an example, assume that a 1200-pound vehicle is boosted to 6.97 mps near the surface of the Earth. Let ΔV_n be supplied by a chemical rocket which is part of the 1200 pounds. After burnout, the rocket is dropped and the remainder is designated as injection payload (IPL). For the photon-powered vehicle, the injection payload includes the weight of the sail.

A comparison of the times of flight for a photon-powered vehicle and a ballistic vehicle, injected into their respective heliocentric orbits with the same initial conditions, is also shown in Table III. The times of flight are longer for the solar-powered vehicle, and the useful payload is less since the sail weight is part of the injection payload.

Assuming a sail density of 2×10^{-4} lb/ft², the weight of the sail is listed in Table III. While these weights are not excessive, it can be seen that the controllable sail-area may limit the size of solar sail vehicles.

Solar Sail Work Compared to Electrical Engine Work

Assume a hypothetical electric engine which has a power output capability of 0.1 kw per second per pound of engine weight, and an efficiency of 70 per cent. After a determination of the optimum specific impulse for a particular mission is made, the mass ratio, power, and finally, the engine weight may be computed.

With the above engine set as a standard, a comparison between a solar sail and electrically powered vehicles is shown in Table IV. Here the sail-powered vehicle shows a payload weight advantage. It thus appears that in some instances the solar sail-powered vehicle is worthy of more consideration.

TABLE IV

Comparison of an Electric Powered Vehicle with a Solar Sail Vehicle

T_f (days)	Optimized specific impulse (seconds)	Injection payload (pounds)	Propellant for electric powered vehicle (pounds)	Solar sail weight (pounds)
150	6500	542	249	97
200	7500	822	300	100

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Interplanetary Navigation*

G. M. CLEMENCE†

Summary—**The similarities and differences between interplanetary navigation and navigation on the earth are discussed. It is shown that in interplanetary space, position-fixing can be accomplished by a simple extension of the techniques already familiar to all navigators; but, when very close to the destination, other techniques are more suitable.**

INTRODUCTION

By interplanetary navigation is meant the art of getting a ship from one planet to another, particularly from the earth to the nearest planets, Venus and Mars, and back again. In many respects it resembles navigation on the earth. There are the same three phases: departure, midcourse, and arrival, each with its separate problems, and the navigator is interested in the same things, namely, his position, course, speed, and time.

There are also some important differences between interplanetary navigation and earthbound navigation. Most important of them is the force of gravity, which dominates all other considerations in all three phases of a voyage. At departure and arrival, considerable power is needed to overcome gravity; in midcourse, gravity is an important aid. Indeed, with proper planning, gravity can be made to provide all the transport required in midcourse, shipboard power being needed only for small changes in course, and for attitude control, thus achieving maximum economy of power at considerable sacrifice of time. For example, a voyage to Mars or Venus in free flight would require about half a year, which could be considerably reduced by using shipboard power for acceleration during the first half of the journey and deceleration for the remaining half.

Another important distinction is that whereas earthbound navigation usually involves a man on board, interplanetary navigation does not, at least after departure. It is beyond the capacity of any man, even with the assistance of a good calculating machine, to make the necessary observations, calculations, and decisions that are necessary for safe arrival, in the short time that is available for making them. The navigation system and control must be completely automatic. If a man is aboard, it is conceivable that he might be permitted to make a limited number of very general decisions, such as, for example, to circle his destination instead of landing, or to return home before reaching his destination, but any decision requiring much technical reasoning must be avoided, because it will not be practicable for him to acquire the necessary experience in advance of his first voyage.

THE GRAVITATIONAL FIELD

The gravitational field of the solar system is difficult (or impossible) to visualize accurately, but a rough two-dimensional analog of it may be easily conceived. Imagine a large circular sheet of thin elastic material stretched over a rigid horizontal hoop. Place a heavy weight representing the sun in the center of the sheet, which will then take a bowl-like shape; the shape of the plastic sheet represents the gravitational field of the sun. A small object placed anywhere on the sheet will roll at once into the sun, somewhat as a body would move in the actual solar system if released almost anywhere in it. Now take a marble representing Jupiter, 1/1000 as massive as the sun, and make it roll in a circle around the inside of the bowl, at such a speed that it neither spirals in to the sun nor climbs up to the edge of the bowl. The actual orbits of the planets differ slightly from circles, but not perceptibly so on a scale that could be reproduced for our model. In the model, the depression in the sheet caused by the weight of Jupiter represents the gravitational field of Jupiter, and must be thought of as extending over the entire sheet, quite deep at Jupiter itself and very shallow at moderate distances, but nevertheless perceptible all over the bowl. A small object now dropped on the sheet will usually roll into the sun as before, but if it happens to fall sufficiently close to Jupiter it will roll into Jupiter, or, if given the proper distance and velocity, may continue to roll around Jupiter, becoming a satellite. Jupiter goes once around the sun in 12 years.

One-fifth of the distance from the sun to Jupiter is the earth, 1/300 as massive as Jupiter, circling the sun once a year, with its gravitational field superposed on those of Jupiter and the sun. A little inside of the earth is Venus, of about the same size, and a little outside is Mars, about 1/10 as massive as the earth. Venus goes around the sun in 225 days and Mars in 687. Beyond Jupiter are Saturn and other planets, and inside of Venus is Mercury.

The three stages of the navigational problem on the sheet, limited to gravity alone, are as follows. At departure it is necessary to give a sufficient velocity to the ship, about 7 miles per second, to make it escape from the depression caused by the earth, and such a direction that it will take its own orbit around the sun, coming as near as may be to, say, Mars during its first trip around. A direct hit is not to be expected, because under the most favorable conditions the accuracy of aim required is much like aiming at a dime at a distance of 160 yards. Therefore, some midcourse guidance will be required. As Mars is closely approached, some deceleration will be necessary for a soft landing.

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CHARTING THE SOLAR SYSTEM

A chart of the solar system is as necessary for interplanetary navigation as one of the surface of the earth is for earthbound navigation. The solar system must be charted in three dimensions instead of two, and since the bodies in the system are all in rapid motion relative to one another, the chart takes an unconventional form. It consists of a timetable or *ephemeris* showing the coordinates of the sun and planets at such short intervals of time that they may be gotten for any instant by interpolation. These tables, together with a short list of the coordinates of a few dozen bright stars, suffice for all purposes. The construction of such tables has been one of the principal duties of the Naval Observatory for more than a century. They are presently available for some of the planets a century into the future, and for all the bodies of interest to the year 1972, with adequate precision for any interplanetary voyage.

Probably the most convenient system of coordinates to use is one of ordinary rectangular coordinates centered in the sun, with the astronomical unit as the unit of distance. The astronomical unit is very nearly the distance from the earth to the sun (about 90,000,000 miles) and it is more convenient than any other for interplanetary distances. In such a system, the actual coordinates of the stars would not be given. The stars are so remote that from any place in the solar system they may be regarded as fixed beacons in the sky, and their direction cosines are more convenient than the actual coordinates would be.

DEPARTURE

The departure of an interplanetary ship is in all respects similar to that of a multistage rocket with two exceptions. First, a great deal of calculation must be done in advance to determine the optimum orbit, which will depend very considerably on whether the midcourse is to be traversed under power or not, and which, in early voyages at least, will be planned so that the ship will travel the shortest practicable distance. The problem is one of compromising between power consumption and time required for the voyage, and a very great range of ways of making the compromise is possible. Having once decided on the orbit, the second exception appears: there is in general only one possible instant of time of departure when the conditions can be fulfilled. Of course, for times of departure very near the optimum one, only small increases of power and/or small increases in the duration of the voyage will be necessary. Probably a moderate compromise in a particular case would be one requiring variations of a few per cent in quantity of fuel and duration of the voyage. A typical departure schedule might read, depart between 22 and 23 hours on any night during the week of August 24, 1960. In such a case, provision would have to be made for varying the direction of departure by a few degrees depending on the date and on the time of day.

MIDCOURSE NAVIGATION

In order to reach Venus or Mars in free flight, it would be necessary to control the direction at departure to a few seconds of arc, and the velocity to comparable precision, say a part in a hundred thousand. Since it is not to be expected that this precision will be attained, it will be necessary to determine the position, course and speed, and to make alterations of course and probably of speed during the voyage. The more frequently these alterations are made, the smaller they need be, and the less total energy will be required for such purposes, up to the point where the alterations become so frequent as to be random in character.

Translated into astronomical terms, the determination of position, course and speed is equivalent to the determination of the orbit, and of the position of the ship in it. Broadly speaking, there are two possible techniques. The first is Doppler radar and ranging, which is best done from the earth, with a transponder in the ship, and is most useful during the first few days of a voyage, when the motion of the ship is principally away from the earth, that is, principally along the line of sight. The technique becomes useless when the ship is moving nearly at right angles to the direction of motion of the earth, as it will be during later stages of the voyage. In this case it is necessary to employ angular measurements, which can best be made from the ship itself, and which should be used at all stages of a voyage, provided the ship is not too close (say 100,000 miles) to the earth or to its destination.

Utilizing angular measurements from the ship, there is no reason why a continuous indication of position should not be obtained by completely automatic means, using trackers and computers within the ship, and communicating the results both to those on board and back to the earth. Course and speed can be obtained with all necessary precision by comparing positions obtained at different times, and these also can be calculated automatically.

The principle of position-finding in interplanetary space is as follows. Suppose that a planet is observed at some instant in coincidence with a star. It follows that the ship must be somewhere on a line joining the star and planet. If another planet is observed simultaneously in coincidence with another star, another line is established intersecting the first one, and the ship's position is at the intersection. In practice the problem is not quite so simple, because it is not convenient to observe coincidences of planets with stars, but the principle is easily generalized. What is done in practice is to measure the angle between a planet and a star in the same general region of the sky. The star is so far away that it may be considered to be infinitely distant. The observation then locates the ship on the surface of a cone having its vertex at the planet and its angle equal to twice the observed angle. A similar observation of another star and planet locates the ship on another cone, which in general intersects the first one in a curved line. In order to fix the position of the ship along this line, a third similar observation is needed (see Fig. 1). It is not necessary, however, to observe three distinct planets; two will do if a suitable choice of stars is made. Also the sun can be used in place of a planet. The planets that are nearer to the ship give more precise indications than distant ones do, and within the inner reaches of the solar system the most valuable objects will be Mars, Venus, the earth, and the sun, although Jupiter and Saturn may be used on occasion. It turns out that there are just enough suitable planets in the system to make the technique a convenient one.

For position-finding by angular measurements, it is not necessary to have a clock on board. A fourth observation of the same type as the other three suffices to determine the time with sufficient precision, since the four cones of position can intersect in a single point at only one instant of time. It will, however, probably be convenient to carry a clock, which need not be a very precise one; a clock keeping time to a few seconds a year is easily provided and is amply precise.

The position-fixing will be very rough by earthbound standards. In the middle of the voyage, errors of 20,000 miles will not be uncommon, but such errors will have no more serious consequences than errors of a mile do on the earth. They correspond to changes in direction of about a minute of arc, which is a convenient order of magnitude for effecting changes in course.

Knowing the position, course and speed, a calculation made on board will show whether the ship and its intended destination are traveling on courses that intersect or not, and whether they will arrive at the intersection at the same time or not. If not, further calculation will be required to show whether the course or the speed, or both, should be changed, and by how much. The calculations are elaborate ones using many significant figures, and cannot possibly be made in the required time without a good stored-program electronic calculator.

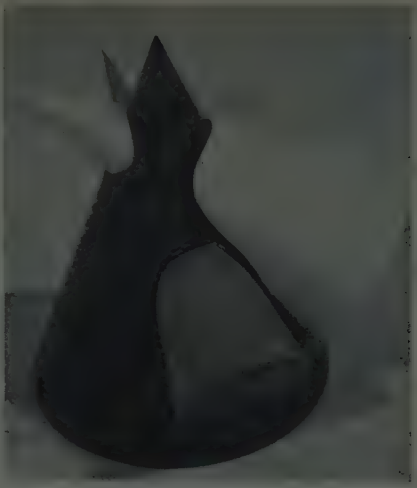

Fig. 1—Three cones of position showing (in black) lines intersecting at position of ship. (Official U. S. Navy photograph.)

ARRIVAL

As the destination is approached within say 100,000 miles, a different technique of navigation must be employed. Although angular measurements are ideal for position-fixing at great distances, it is not possible to rely on them when landing. At this stage, the distance and velocity of the ship relative to the landing field must be known with high precision, and it will be desirable to employ radar ranging and Doppler radar, using signals transmitted from the ship. It will be necessary to take into account the relative orbital velocity of the destination, and its rotational velocity as well, unless a landing is to be made at one of the poles. The rotational velocity of Mars is known already with enough precision, but the same may not be true for Venus. Venus is perpetually covered with dense clouds, and no permanent markings on the surface have ever been observed with certainty. Perhaps special observations at short range will be necessary for determining its rotational velocity.

Navigation Using Signals from High-Altitude Satellites*

ALTON B MOODY†

Summary—**If electronics can be satisfactorily applied to celestial navigation, a system having universal coverage without weather limitations might be produced. Radio stars are not a promising possibility, but the sun and moon are being tracked by the U. S. Navy Radio Sextant, AN/SRN-4, to provide limited coverage. Universal coverage might be achieved by the addition of artificial earth satellites. Satellites in orbits a few hundred miles from the earth might be used in some form of piloting system, but orbit-prediction problems, limited coverage by individual satellites, and computer complexities are serious obstacles to be overcome by such a system. A different approach would be to place three or four satellites in orbits at an optimum distance somewhere between 1000 and 12,000 miles from the earth to serve as artificial celestial bodies in a system that would be a natural evolution from traditional celestial navigation methods. A stabilized directional antenna operating with a receiver capable of accepting signals from both the satellites and the sun would provide angle measurement data both for fixing the position of the craft and also for establishing a north reference. The degree of sophistication of the user equipment would differ with requirements.**

INTRODUCTION

NOW that the novelty of artificial satellites of the earth has worn off, the question is logically being raised as to their practical utility for purposes other than pure research. Natural celestial bodies which have been available throughout recorded history may provide some guidance. These bodies, in addition to providing light and heat and serving certain other useful purposes, have been used for navigation to assist travelers on land, at sea, and in the air to fix their positions and determine direction.

Celestial navigation has been attractive because it provides universal coverage over the entire world with essentially uniform accuracy adequate to meet the requirements of the navigator. Its principal disadvantage has been its dependence upon clear weather. Other shortcomings have been the lack of a fully satisfactory horizontal reference for making the observations, and the need for lengthy computations to reduce the observations to positional information related to the observer.

Electronics has aided materially in the solution of these last two problems, but what of the larger problem of weather limitations? Here, too, electronics has had a significant role. Radar, loran, radio ranges of various types, Decca, the radio direction finder, and other electronic piloting aids have contributed significantly to the removal of the weather limitation. However, to a greater or lesser extent these aids are limited by geographical siting problems, propagation patterns, logistic considerations, vagaries of propagation over long paths within the atmosphere, ionospheric effects, the need for international cooperation, diplomatic considerations, antenna destruction by weather or future enemy action, the threat of possible countermeasures, lack of military security, and high installation and maintenance costs. Electronics has also been used to produce two excellent types of dead reckoning systems, Doppler radar and inertial, but these are not wholly adequate because, being dead reckoning systems, their results degrade with time and distance traveled.

Can celestial navigation and electronics be combined to produce a system having the universality of celestial and the all-weather capability of electronics, with essential freedom from the limitations encountered by present systems? This is a subject worthy of consideration.

ELECTRONIC CELESTIAL NAVIGATION USING NATURAL CELESTIAL BODIES

Discrete sources of cosmic radio energy, commonly called "radio stars," have been studied with this in mind.[1] Angle measurement would be utilized in the traditional manner of celestial navigation. However, the signal received from known natural sources is weak, and decreases rapidly as wavelength is decreased. As wavelength increases, the angular sensitivity of a tracking system decreases. Studies taking into account various considerations such as receivers, antennas, source characteristics, atmospheric attenuation, and refraction indicate an optimum wavelength region centered at about 4 cm. For a practical navigation system of suitable accuracy an antenna diameter of more than 15 feet would probably be necessary to overcome atmospheric emission effects. A considerable amount of research would be needed to accomplish this. Eventually such a system might evolve, but radio stars are not promising as a means of providing a satisfactory electronic celestial navigation system in the near future.

Both the sun and moon have been satisfactorily tracked by means of the Radio Sextant, AN/SRN-4, developed for the U. S. Navy by the Collins Radio Company. The installation aboard the *USS Compass Island* is shown in Fig. 1. Two optimum wavelength regions have been identified. At 8.7 mm the angular sensitivity of an antenna of given size is relatively great, and solar dynamic activity is not troublesome, but the

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[1] G. R. Marner, "Automatic radio-celestial navigation," *J. Brit. Inst. Navigation*, vol. 12, pp. 249-259; July/October, 1959.

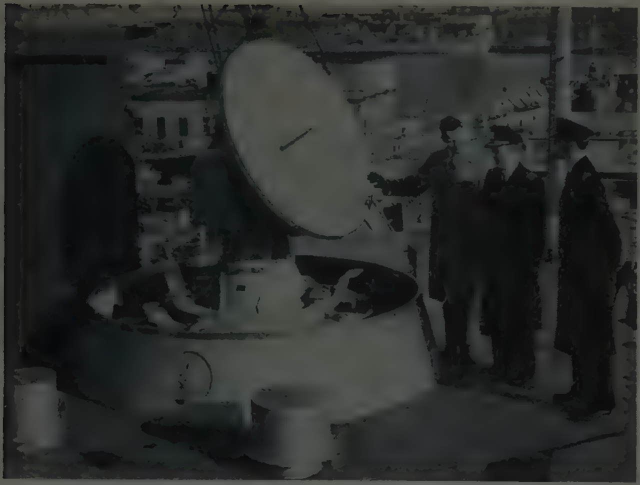

Fig. 1—Radio Sextant AN/SRN-4 (courtesy of Collins Radio Company). (Official U. S. Navy photograph.)

operation is adversely affected by rainfall. At a wavelength of 1.9 cm the system is affected less by severe rainfall, and an antenna only slightly larger than at 8.7 mm is needed. At wavelength longer than 2.0 cm, solar dynamic activity effects are great enough to be troublesome. Although the moon has been successfully tracked, as indicated, the inclusion of this capability adds considerably to the complexity of the system.

Although the sun and moon have been used successfully by the Navy, these bodies are not always available. The length of the periods when neither is available increases as latitude increases, becoming maximum at the poles, where intervals as long as two weeks occur during which neither body is above the horizon.

It is appropriate, therefore, that the possibility of using artificial earth satellites be explored to determine the practicability of using a modest number of them to provide a system that will meet navigational requirements at a price the user will be able and willing to pay.

ELECTRONIC CELESTIAL NAVIGATION USING ARTIFICIAL EARTH SATELLITES

The ability to place satellites in orbit and receive signals from them over a long period of time has been amply demonstrated. The experience with Sputnik III and Vanguard I are notable examples. There remain to be explored, however, the requirements for a navigational satellite, and the problem as to whether or not these requirements can be met at reasonable cost.

First, what are the parameters that might be measured? The possibilities can be summarized as

1) Distance and its time derivatives,
2) Altitude and its time derivatives,
3) Azimuth and its time derivatives.

The measurement of distance and its time derivatives is essentially a piloting problem, the position of the observer being established with respect to the satellite. Further, if the distance at nearest point of approach is determined, direction is also established and one obtains a fix. A fix might also be obtained by means of simultaneous distances from two or more satellites, or a running fix by two distances of the same satellite after a satisfactory change of azimuth.

One method of determining distance would be by direct measurement using primary radar. From a military standpoint this is not attractive because of the need for transmission by the user. The same problem would arise with respect to secondary radar using a transponder in the satellite. Additionally, a saturation problem might be involved if a large number of users interrogated the satellite at the same time. Jamming by a future enemy would not be difficult.

A more attractive method of determining distance would be by measurement of the Doppler shift in frequency of CW transmissions from the satellite. This method has been used successfully in the United States and the USSR to determine the positions of satellites from known positions on earth. The Doppler frequency change at any moment could be determined only if one knew the exact frequency of transmission, but if the variations of the Doppler effect throughout a passage of the satellite were measured and analyzed, knowledge of the exact frequency of transmission would not be needed, as long as it remained reasonably stable during the period of observation.

Analysis of the Doppler information to extract useful navigational information would require computational facilities of considerable complexity. Use of Doppler favors orbits relatively close to the earth. As the radii of orbits increase, the Doppler effect becomes smaller and less sensitive to the position of the observer. Also, unless stabilization or tracking is used, the antenna would have to be reasonably omnidirectional. This would favor relatively long wavelengths if adequate collecting area were to be provided with reasonable power. As wavelength increases, unpredictable ionospheric phase shifts and atmospheric refraction become a greater probability. If multiple frequencies are used to provide corrections, added complexities are encountered. In order to prove the feasibility of a satellite navigation system using Doppler techniques, it will be necessary to indicate solutions in the areas mentioned above and to demonstrate ability to predict orbits with an accuracy considerably better than attained to date. This problem is discussed in more detail later. Studies along these lines are presently being pursued under a U. S. Navy contract.

A completely different approach to the possible use of satellites for navigation, presently under study, is to consider them as celestial bodies and to use the traditional angle-measurement approach of optical celestial navigation. As commonly practiced, this involves measurement of the altitude (angular distance above the horizontal) of two or more bodies simultaneously, or of one body with a sufficient time separation of observations to allow for an adequate change of azimuth. Each observation establishes a circle of position with the geographical position of the body (the terrestrial position at which the celestial body is vertically overhead) as the center and the zenith distance as radius. A small portion of this circle in the vicinity of the observer is called the "Sumner Line," or line of position. The common intersection of two such lines fixes the position of the observer.

At the distances of natural celestial bodies, their change of apparent position is essentially that caused by daily rotation of the earth upon its axis. For moderate altitudes, this requires the availability of at least two celestial bodies for wholly satisfactory results. With a single body, a wait of several hours between observations permits introduction of unpredictable dead reckoning errors that can be intolerably large. At the faster rates of satellites, the running-fix principle is more attractive.

A single body might provide a celestial fix by simultaneous measurement of any two of the quantities, 1) altitude, 2) rate of change of altitude, 3) azimuth, and 4) rate of change of azimuth. Methods involving azimuth and rate of change of azimuth have been virtually unused because this information has not been available to the required accuracy. Likewise, rate of change of altitude of natural celestial bodies involves advanced averaging techniques. However, with the higher rates involved with artificial satellites, this parameter is worthy of consideration. One feature of using altitude and rate of change of altitude that is particularly attractive is that the two sets of curves involved are always orthogonal to each other, which is the ideal situation in celestial navigation. Whether rate of change of altitude can be determined satisfactorily without undue complexity of equipment has not been established.

It is well established, however, that navigation using simultaneous altitudes of two bodies, or successive altitudes of the same body, can provide a satisfactory navigation system without complicated averaging processes. Further, this method lends itself to hand computation methods which are well understood by navigators everywhere. The method is commonly used daily with natural celestial bodies. The use of satellites would require adequate electronic equipment for measurement of the parameters involved. No analysis of results would be needed. The computation required to convert the measured parameters to position of the observer, or to provide guidance, could be performed electronically where this is desirable without adding greatly to the complexity of the equipment.

Although details relating to exact procedures and specifications for almanacs and sight reduction tables have not been worked out, these are problems which differ in no important respects from those which have already been adequately solved for celestial navigation using natural celestial bodies.

This method, then, is a logical evolution of the celestial navigation commonly practiced today, and it involves no untried principles. Its success depends primarily upon the ability to predict future positions of satellites with sufficient accuracy, and since the size of the orbit is not critical, an optimum value can be found. The orbit problem is discussed in more detail later.

If angle measurement is to be used in a satellite navigation system, stabilization and a directional antenna will be needed. This requirement adds complexity and cost to the equipment, but poses no serious technical problems. If short microwaves are used, the Radio Sextant, AN/SRN-4, can be used with relatively minor modification. This equipment, already developed, includes satisfactory stabilization and a directional antenna. The use of this equipment would obviate the need for long feasibility studies and time-consuming developmental work. It would also permit utilization of the sun, and possibly the moon, to extend coverage and provide a system of limited availability if artificial satellites were to be eliminated either by natural causes, internal failure, or possible enemy countermeasures. The success of the Radio Sextant is due largely to the efforts of Drs. D. O. McCoy and Gene Marner, both of Collins Radio Company.

Stabilization and a directional antenna make possible a highly important north-reference capability not available in other methods. They also reduce the complexity of equipment needed in the satellites; this is an important consideration in view of their inaccessibility for service in the event of malfunction or failure.

ORBITS

The type of satellite navigation system selected is closely related to the orbit problem; this may well be found to be the controlling consideration. Questions of orbital predictability, coverage, and number of satellites required are basic considerations.

The predictability of satellites launched to date has not approached the requirements for navigation. The perigees of available satellites have all been within a few hundred miles from the surface of the earth. The nature and magnitude of the disturbances to such orbits have not been fully defined. It is well known, however, that unpredictable effects caused by lack of precise knowledge of the gravitational field of the earth and variable atmospheric effects have posed serious problems. Astronomers who have been close to the prediction problem of satellites launched to date are not optimistic of an early solution.

As orbits increase in size, the disturbing gravitational and atmospheric effects become progressively less pronounced. However, as orbit radius increases, perturbations caused by gravitational attractions by the various natural celestial bodies increase. At very large radii these disturbances become excessive and difficult to predict. This would indicate that there is some intermediate radius which will prove optimum from a prediction standpoint. This radius has not been established, but it will undoubtedly be measured in thousands of miles, in addition to the radius of the earth. This is in sharp contrast to the orbits of satellites launched to date, all of which have had perigees relatively close to the earth. When compared with these, the satellites in the optimum orbit can be considered "high altitude." These are the satellites anticipated in the title of this article.

An important consideration in determining the number of satellites needed is the coverage provided by one such body. At a distance of 400 miles from the surface of the earth, a satellite can be seen simultaneously over less than 5 per cent of the surface of the earth. This increases to about 10 per cent at a height of 1000 miles, 25 per cent at 4000 miles, about 35.5 per cent at 10,000 miles, and about 50 per cent at infinity. Fig. 2 shows the percentage of the earth's surface over which a satellite is observable at one time.

Another approach to the coverage problem is to consider the width of the belt in which a satellite can be observed during a passage around the earth. At 400 miles from the earth this is about 45°, at 1000 miles it is about 88°, at 4,000 miles about 120°, at 10,000 miles about 143.5°, and at infinity about 180°.

Where some form of distance measurement is involved, the belt within which the satellite can be effectively used for navigation might be somewhat less than the radio visible belt because of limitations in the distance at which distance measurement provides a satisfactory parameter.

If the running-fix principle is to be employed with a series of observations of a single body, it is desirable to consider the width of the belt in which the azimuth changes by some established number of degrees during a transit. If 90° is used as the minimum change in azimuth (90° between observations being the optimum), the belt is about 36° at 400 miles from the surface of the earth, 57° at 1000 miles, 101° at 4000 miles, and 134° at 10,000 miles. During one revolution of the satellite, the earth would rotate on its axis about 25° for a satellite at 400 miles, 30° for one at 1000 miles, 60° at 4000 miles, and 140° at 10,000 miles. Fig. 3 shows the widths of the radio belt, width of the belt in which azimuth changes more than 90°, and earth rotation during one orbital period for various orbits.

The faster motion of a satellite at low altitudes is an advantage to any method involving a time derivative, but it also imposes a more stringent time requirement. Thus, the time tolerance for an error of $0'.1$ in subsatellite position is less than 0.03 second for a satellite 400 miles from the earth. This increases to nearly 0.035 second at 1000 miles, 0.067 second at 4000 miles, and 0.155 second at 10,000 miles. An error of $0'.1$ in orbit prediction may be too large for a high accuracy navigation system.

An error in the height prediction introduces a parallax error which is significant where angle measurement is involved. Parallax error is maximum at the horizon and decreases to zero at the zenith. If a parallax error of $0'.1$ can be accepted, the tolerance in height prediction is less than half a mile for a satellite at a distance of 400 miles from the surface of the earth, a little more than half a mile at 1000 miles, about 1.3 miles at 4000 miles, and

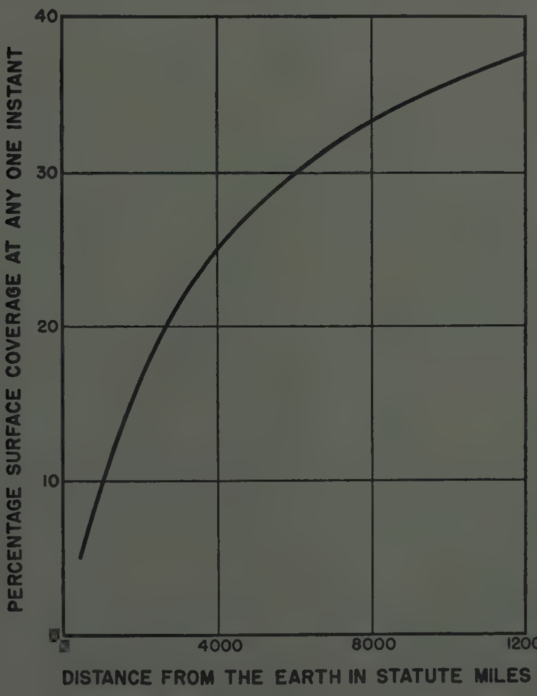

Fig. 2—Percentage of earth's surface in which a satellite can be seen at one time.

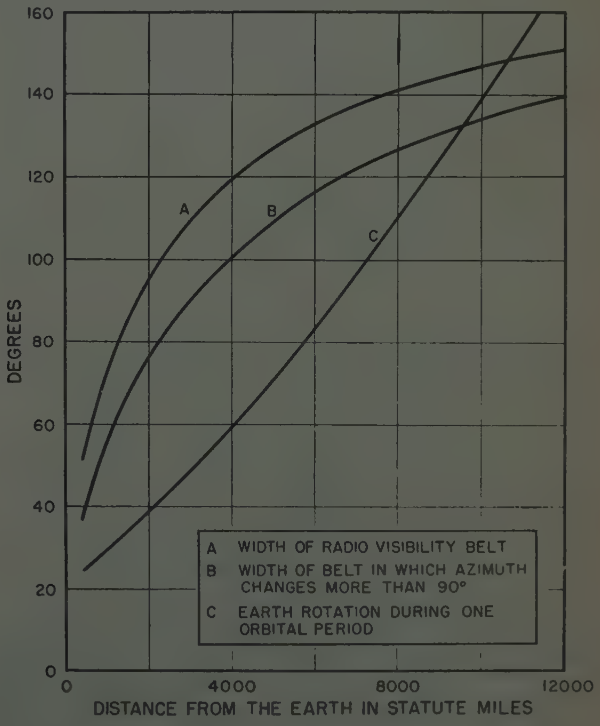

Fig. 3—Width of visibility belts and earth rotation during one orbital period.

a little more than 4.1 miles at 10,000 miles.

Thus, it appears evident that the problems of predicting future positions of satellites and providing adequate ephemeridal information are greater for satellites relatively close to the earth than at a distance of several thousand miles. Not only are small orbits most difficult to predict, but the accuracy requirements are greatest for these orbits.

The number of satellites required for a navigation system depends upon the parameters to be measured, the distance of the satellites from the earth, interrelationship of the orbits and the satellites in their orbits, and the duration and frequency of the intervals during which no fix is available. It appears evident from the geometrical considerations discussed earlier that for any type system, the number of satellites needed if large periods of unavailability are to be avoided is greater at orbits close to the earth then for those at a distance of several thousand miles.

If the system requires a relatively long period of observation to establish a fix, the number of satellites will be greater for a given fix frequency. The maximum needed would be for a system requiring satellites relatively near the earth and a full pass of a satellite to establish a fix. The minimum would be required for a system using high-altitude satellites and providing continuous fixing capability during passage of a satellite above the horizon.

The number of satellites needed for any system can be kept to a minimum by judicious selection of the orbits and the positioning of the bodies within the orbits. However, the value of this device is somewhat limited by the lack of capability to place a satellite precisely in a predetermined orbit. Because of differences in orbits, variations in precession and rotation of the orbits can be expected to nullify any planned relationship of position of satellites with respect to each other. The one element of the orbit that can be controlled with sufficient accuracy is inclination to the equator. It appears from a preliminary study that a family of orbits at progressive angles of inclination will best serve the needs of any type system.

If infrequent periods of nonavailability of several minutes is acceptable, with occasional periods of more than an hour, a preliminary investigation indicates that a high-altitude system with three to four satellites can provide satisfactory coverage. The coverage can be further improved if the system can also utilize natural celestial bodies. If the ability to reduce differences in the launching of satellites in predetermined orbits is demonstrated, the number of satellites can be decreased.

Any statement of the number of satellites needed for a given system should be evaluated in terms of the coverage it provides and its dependence upon capabilities theoretically possible but not yet demonstrated.

One final consideration relating to the size of the orbits concerns the tracking to provide ephemeridal data. Both optical and electronic tracking are probably desirable. The required number of tracking stations decreases with the distance of the satellite from the earth because of the greater area over which such satellites can be seen. At distances of only a few hundred miles from the surface of the earth, and also at distances at several tens of thousands of miles from the earth, the unpredictable nature of the perturbations would require that a closer watch be kept on the satellites.

Thus, all factors favor an orbit that avoids both extremes of distance from the earth. While the limits of the desirable range of orbits have not been definitely established, preliminary study from available data seems to indicate that the optimum distance lies somewhere between about 1000 miles and 12,000 miles from the surface of the earth. Use of this range would make possible the development of a system utilizing short microwaves so that the sun and possibly the moon might be used with the satellites to increase coverage. Such a system would be a natural evolution—actually a relatively short step—from systems already well developed.

The use of satellites in orbits thousands of miles from the earth appears to preclude the use of distance and its derivatives as the primary method of navigation by satellites. However, there are so many areas of uncertainty yet to be resolved if any practical navigation system using satellites is to evolve, that it is premature and unrealistic to pursue any one possibility to the exclusion of others until a clear superiority is demonstrated. The optimum system might even prove to be one providing both piloting and celestial capabilities, perhaps with somewhat overlapping ranges of application.

THE HIGH-ALTITUDE SATELLITE

What are the requirements for a high-altitude satellite to meet the needs of a navigation system based upon angle measurement and possibly its time derivatives, and providing high-accuracy north reference?

Perhaps the first consideration is one of wavelength. It has been pointed out previously that use of short microwaves will permit use of the sun and possibly the moon to augment the coverage of artificial earth satellites and to provide partial coverage if satellites should become unavailable. However, considerations of the satellite system itself favor a short microwave frequency to provide adequate directivity in the user's receiving antenna, without an unsatisfactorily large antenna. The use of short microwaves increases the transmitter power requirements over what they would be in the UHF and VHF region, but these requirements are not excessive.

The use of short microwaves avoids the major propagation difficulties related to frequency instability and refraction as the signals pass through the ionosphere and lower atmosphere of the earth. Studies in this area carried out on Navy contracts have yielded sufficient information to conclude that this will not be a problem, as it can be expected to be if longer wavelengths are used.

The function of the transmitter in the satellite would be to provide a CW signal of adequate power for tracking. Since it has no other function, it can be of relatively simple design, enhancing the probability of a long, useful life.

Batteries, solar cells, fuel cells, and thermionic converters are possible power sources, but a combination of solar cells and batteries appears most attractive at this time. This is a proven combination. If the protective covering of the solar cells can be omitted, as seems probable, a considerable saving in weight can be effected. Some form of nuclear power may eventually be available for use in satellites.

If the orbit were exactly circular, and the satellite always maintained the same orientation with respect to the earth, the transmitting antenna could be shaped so that all users within the entire area of illumination on the earth would receive the signal with equal intensity. To achieve this would require a launching capability beyond that anticipated in the near future, and satellite stabilization about at least two axes. Even if successful stabilization is achieved, its use would add complexity and increase the probability of failure. A realistic approach is the use of an antenna with a pattern as nearly omnidirectional as possible. Circular polarization is desirable to avoid polarization error. The design of a suitable antenna is reasonably straightforward and no major problem would be anticipated in this area.

The weight of the transmitter, with adequate protection against vibration and shock during launching, and with suitable provision for removal of heat from the satellite during operation, should be within the capability of available launching devices and probably not more than 30 to 50 pounds.

User Equipment

The size and complexity of the user equipment for a celestial navigation system based upon CW signals from high-altitude satellites would depend upon the parameters to be measured, the acceptable period of time between fixes, the accuracy requirements, and the degree of sophistication desired in the readout. One feature of the use of relatively simple high-altitude satellites is that a wide choice of user equipment capability is possible with the same signals. This might vary from relatively simple equipment providing altitude and azimuth of moderate accuracy for hand computation, as with traditional celestial navigation, to highly complex equipment providing continuous position information or actual guidance according to a predetermined schedule. There is no reason to doubt that accuracy requirements can ultimately be achieved by means of such a system.

The first equipment could logically be expected to be an evolutionary development of the Radio Sextant, AN/SRN-4. There are two major components of this instrument. One is the tracking unit, consisting of the antenna and its scanning mechanism, the receiver, the tracking servos, and the angle readout system. The other major component is the Schuler-tuned stabilization system and north reference.

The properties of the receiving system, of course, are closely related to the transmitter power requirements. In satellite tracking, the transmitted power requirement can be kept to a minimum by the use of very narrow bandwidth. The phase-lock type receiver, first developed by the Jet Propulsion Laboratory for missile tracking purposes, offers a promising solution. In contrast, sun tracking involves very wide RF bandwidth. Consequently, a receiver designed for tracking both artificial satellites and the sun should have the capability of operating in both the narrow-bandwidth phase-lock mode and the wide-bandwidth radiometric mode. A number of components could be common to both modes of operation. Since the phase-lock frequency tracking loop is of optimum design to phase-track an incoming signal accurately, it provides the optimum method of extracting the Doppler signal, if this proves desirable.

A time standard of such accuracy that only occasional resetting by means of time standard broadcasts appears feasible.

The readout system to provide altitude and azimuth would be designed in accordance with accuracy requirements of the individual system.

As pointed out earlier, the complexity of the user equipment can vary between wide limits to meet different requirements. The simplest equipment would provide only altitude and azimuth for use in any manner that such information obtained optically is used. To this might be added a suitable computer with the capability of determining a fix by the running-fix principle, simultaneous observation of two or more bodies, or by altitude and rate of change of altitude. During the relatively brief and infrequent periods when a fix would not be available, the dead reckoning might be maintained by hand plot, hand computation, mechanical dead reckoning equipment, or by a more sophisticated system such as Doppler or inertial, according to requirements. Many of the elements of a high-grade inertial system would already be available within the receiving and stabilization system. The ultimate system would have the added capability of providing guidance in accordance with a previously inserted track and time schedule.

The system discussed would be suitable for use aboard a surface ship. It could also be used by a submarine, with the requirement that a parabolic antenna of moderate size be exposed periodically above the surface for brief intervals to obtain a fix. Between fixes, positional information would be provided by a high-grade, built-in inertial system. Aircraft and even missiles might be able to use the system, perhaps with some loss of accuracy to meet space and weight requirements.

CONTRIBUTIONS OF HIGH-ALTITUDE SATELLITE SYSTEM

Certain contributions might be expected from a high-altitude satellite navigation system of the type discussed. These are summarized as follows:

1) Continuous all weather position-fixing capability with universal coverage of uniform accuracy.

2) A continuous north reference in all weather.

3) Great flexibility in user equipment to meet different requirements.

4) High potential accuracy.

5) A passive system without ground-based transmitter problems. No transmission of any kind would be required from the surface of the earth.

6) A virtually jam-proof system.

7) A system combining use of artificial earth satellites with at least one natural celestial body to provide flexibility, extend coverage, and reduce the possibility of successful countermeasures by a future enemy.

8) Maximum utilization of satellites by using orbits at such distance that a minimum number would be needed, and these having a small number of relatively simple, reliable components to reduce obsolescence and increase useful life.

9) Limited ground facilities required. Because of the high, stable orbits, the only ground facilities required would be launching equipment, one radio and perhaps three optical tracking stations, and orbital computation and dissemination facilities. None of these would be of great complexity with the possible exception of the radio tracking station, and a suitable one is already in existence in the U. S. Navy-owned Feather Ridge station in Iowa.

10) The system could be provided as a short, orderly evolution of methods and equipment already developed, using principles in common use daily by navigators everywhere.

CONCLUSION

A preliminary study indicates that the development of a celestial navigation system using a small number of artificial earth satellites at an optimum distance expected to be somewhere between 1000 and 12,000 miles from the surface of the earth is feasible, and that such a system could be developed as a logical evolution of traditional celestial navigation and the Radio Sextant, AN/SRN-4, developed for the U. S. Navy for use with the sun and moon. It is entirely possible that the cost of development and maintenance of such a system, particularly if use is made of natural celestial bodies, would be less than that of a ground-transmitter system providing world-wide coverage.

If the need for a navigation system based upon signals from satellites is established, it would appear most logical to utilize the advantages of relatively high-altitude stable orbits and the work already done in developing the Radio Sextant and studying short microwave propagation through the atmosphere, rather than to pursue development of a system of more uncertain future. However, it is premature to rule out other possible methods of utilizing satellites for navigation until a clear superiority is demonstrated.

The opinions or assertions in this paper are those of the author and are not to be construed as official, or reflecting the views of the Department of Defense, the Navy Department, or the naval service at large.

A Satellite Doppler Navigation System*

W. H. GUIER† AND G. C. WEIFFENBACH†

***Summary*—It is shown that a satellite Doppler navigation system can provide navigation to at least one-half mile accuracy from a single pass of a satellite, provided that proper use is made of the full Doppler curve. While not extensively discussed, a satellite Doppler navigation system is presented which conveniently provides navigators with all necessary inputs for position calculation, including up-to-date orbital data. The main body of the paper presents the results of a comprehensive error analysis which proves the feasibility of achieving navigational accuracies of one-half mile using such a system under realistic conditions.**

I. INTRODUCTION

IN late 1957 and early 1958, the authors made a study of the feasibility of tracking satellites utilizing only the Doppler shift of radio signals transmitted from the satellite.[1] These studies indicated that, when properly utilized, each segment of the received Doppler curve provides useful information about the track of the satellite, and that in order to gain the maximum information from the Doppler data, the data should be used in a direct calculation of the six orbit parameters rather than used to compute intermediate parameters such as the slant range and time of closest approach. With this approach, and utilizing all of the Doppler curve for determining the track of a satellite (orbit elements), it was found that a single Doppler curve, received during a single pass of the satellite by a single receiving station, was sufficient to determine the orbit of the satellite to reasonable accuracy.

Table I indicates the results of a determination of the orbit of Sputnik I from a single Doppler curve.[2] In determining this orbit, the theoretical equation for the Doppler shift including ionospheric refraction effects was written as a function of the six orbit elements and two more parameters representing, respectively, a "vernier correction" to the satellite's transmitter frequency and a parameterization of the electron density of the ionosphere. The eight parameters were then varied until the theoretically computed Doppler shift agreed (in the "least squares" sense) with the experimental Doppler shift received from the 20-mc transmitter of Sputnik I. For comparison, the values of the orbit elements determined by two British agencies are given.[3,4] The British orbital data was believed to be the most accurate data available for Sputnik I at that time, and was compiled by the British from many sets of data including both interferometric and Doppler data. For this pass, it can be seen that surprisingly accurate values for all six of the orbit elements were obtained.

TABLE I

ORBITAL DATA
SPUTNIK I—20 MC

Estimated experimental error:	±4 cps
RMS fit to data:	±1.6 cps
Approximate ground range:	73 statute miles
Minimum angle of arrival from horizon:	20°
Time of closest approach:	23:47 GCT, October 21, 1957
Latitude of receiver:	38°41′N
Longitude of receiver:	75°59′W

Orbit Element	Doppler Determination	Reference 3	Reference 4
Period	95 minutes 38 seconds	95 minutes 36 seconds	95 minutes 34 seconds
Eccentricity	0.053	0.053±0.001	0.048±0.002
Inclination	64° 10′	64° 40′±10′	65°
Argument of perigee	43° 30′		
Latitude of perigee	38° 20′ N	36°±3° N	40.9° N
Longitude of ascending node	289°	291.6°±0.3°	

These values for the orbital elements were taken from References 3 and 4 and were given for October 15. Using their values for the secular variations, the values for October 15 were extapolated to October 21 for purposes of comparison.

Clearly, the constraint on satellite tracking that only a single Doppler curve is to be used is not appropriate if the main objective is to determine the orbit of a satellite accurately. This constraint was imposed solely to illustrate the large amount of information that can be obtained from a single Doppler curve. However, if the problem inverse to satellite tracking is considered, namely determining the location of the receiving station assuming a known satellite orbit (navigation), the ability to use a single Doppler curve becomes highly significant, since then a "navigation fix" can be obtained every time a satellite passes within range of the navigator. Furthermore, since navigation requires determining only two parameters (latitude and longitude) instead of the six orbital parameters of the satellite, a very large increase in the accuracy of the measurement can be expected.

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[1] W. H. Guier and G. C. Weiffenbach, "Theoretical Analysis of Doppler Radio Signals from Earth Satellites," Applied Physics Laboratory, The Johns Hopkins University, Silver Spring, Md., Bumblebee Series Rept. No. 276; April, 1958.

[2] "Letters to the Editors," *Nature*, vol. 181, p. 1525; May 31, 1958.

[3] Staff of the Mullard Radio Astronomy Observatory, Cambridge, Eng., *Nature*, vol. 180; 1957.

[4] Staff of the Royal Aircraft Establishment, Farnborough, Eng., *Nature*, vol. 180; 1957.

Such an approach to navigation was suggested in early 1958 by Dr. F. T. McClure. Shortly thereafter, initial studies on the expected accuracy of surface navigation, utilizing the radio Doppler shift in a manner completely analogous to the above example of Doppler tracking, indicated that this method could result in navigational accuracies of better than one-half mile.

In the present report, the main emphasis is placed upon the feasibility of achieving Doppler navigation fixes to an accuracy of one-half mile rather than upon the design and development of such a navigation system from the over-all system point of view. However, for purposes of orientation and background information, Section II contains a brief description of a satellite Doppler navigation system which adequately supplies the proper inputs to the navigator, such as orbital data. The main body of the paper is contained in Section III, which presents the results of various feasibility studies of Doppler navigation within the purview of the system as outlined in Section II. This section includes an analysis of pertinent sources of error including bias errors that are not of a statistical nature. Section IV contains a summary of the results and some comments on the use of such a system for navigation.

II. Elements of the Navigation System

An operational Doppler satellite navigation system should consist of 1) a network of satellites in orbit about the Earth at altitudes optimum from the standpoint of accuracy of both satellite tracking and of navigation, 2) a network of stations supplying tracking data to a computing complex for maintaining accurate ephemerides of the network of satellites, 3) a means for supplying to individual navigators the ephemeris of each satellite, and 4) independent navigating gear designed to receive the Doppler data from the satellite and to determine a navigational fix from these data.

The number of satellites that are required and the orientation of their orbits depend upon the maximum allowable time interval between opportunities for obtaining a fix. One satellite in an approximately circular polar orbit at an altitude of 400 nautical miles provides, anywhere on the Earth, the opportunity to obtain a navigational fix at least twice a day. It can also be shown that four satellites, in circular 400-mile orbits, can provide sufficient coverage for most of the Earth to obtain a navigational fix at intervals of about one and one-half hours. The altitude of the satellites should be sufficiently high to be little affected by air drag, in order to guarantee sufficient lifetime in orbit and to insure the capability of accurate tracking in the presence of the random variations in drag that would be troublesome at lower altitudes. On the other hand, for any given pass of the satellite, the accuracy of the navigation fix decreases with satellite altitude. Furthermore, Doppler data covering all of a satellite pass is required to obtain the maximum accuracy of a navigational fix. The total time that the satellite is visible increases with altitude, thereby increasing the time necessary to get sufficient Doppler data. Satellite altitudes between 400 and 500 nautical miles appear to be about optimum.

Each satellite would radiate very stable CW signals from which the Doppler shift would be measured by the navigator. In Section III it is shown that frequency stabilities of the order of 1 part in 10^8 are required. It is shown that if the Doppler shift is received simultaneously on two frequencies, a continuous real-time correction for ionospheric refraction can be made that does not require prior knowledge of the electron density profile. Frequencies in the 200- to 400-mc region appear to be satisfactory for providing sufficiently accurate refraction corrections and yet are not so high as to preclude the use of transistorized oscillators and transmitters. Finally, it can be shown that a transmitter power of about 100 milliwatts at these frequencies can give ample signal strength on the ground with simple antenna systems (such as a vertical dipole over a ground plane).

It would be highly desirable to provide the ability for each satellite to give the navigator its own ephemerides. Consequently, each satellite should contain a memory which would receive from an injection station on the ground its most recently computed orbital parameters and/or ephemeris. Thereafter, until new orbital data is injected, the satellite orbits around the Earth transmitting on two very stable harmonically related frequencies, and periodically reads out the injected orbital data from its own memory. Besides being a convenient method for identification and a supplier of orbital data to the navigator, the use of a satellite memory allows very frequent updating of the orbit. No matter how accurately the forces acting on the satellite are known, the accuracy of a computed satellite ephemeris decreases with the length of time that it is predicted into the future. With the use of such an injection station—satellite memory system—it becomes feasible to update the orbit of each satellite twice a day, which may be required to maintain satellite ephemerides to an accuracy compatible with one-half-mile navigation accuracy. Furthermore, since the ephemerides need cover only relatively short time intervals, complete minute-by-minute listings of satellite positions can be stored without the memory's weight and size becoming a burdensome factor.

Finally, since the stable oscillator in the satellite can be used as a clock, and the satellite memory can be used to store time corrections to this clock, as well as orbital data, the satellite can also transmit a time-synchronizing pulse and a correction to this time pulse. In this way the navigator (or any receiving station) can be supplied an accurate time standard on a world-wide basis. With the addition of the time standard, the injection station monitors the time pulses from the satellite, and determines a time correction with reference to some accepted standard. The injection station then transmits to the satellite twice each day the latest orbital data received from the computing center and the digitalized time correction. By now, the navigator receives during any single satellite pass all the inputs required for accurate navigation. He receives Doppler data on two harmonically related frequencies, from which refraction-corrected, or "vacuum," Doppler data can be obtained; he receives an accurate time check from which his own frequency and time standards can be calibrated; and finally, he receives the necessary orbital data, from which, together with the Doppler curve, he can compute his position.

Clearly, the accuracy of the satellite orbital data must be compatible with the requirements of navigational accuracies. Since the navigation satellites described above transmit all data necessary to perform Doppler tracking of the satellite, it is natural that the tracking system utilize the Doppler shift for computing the satellite orbit. The authors have reported elsewhere on the ability of a network of Doppler receiving stations to supply data for tracking satellites to an accuracy better than one-half-mile.[1,5,6] Presently it appears that no more than ten tracking stations are needed to supply ample data for updating the orbit every twelve hours, and the degree to which fewer stations can be used is only dependent upon the state of knowledge of the forces acting on the satellites.

Basically, the navigation gear required to receive the satellite signals and compute the navigator's position consists of a dual frequency receiver, a mixer to combine the two received frequencies for elimination of the refraction effects in a manner outlined in Section III, instrumentation to reduce the Doppler data to digital form, and a computer to determine the position of the navigator that provides a best fit between the received Doppler data and the theoretically computed Doppler shift. The sophistication of the computer can vary from the one extreme of hand calculations when lower accuracy is sufficient and time is available, to highly accurate, high-speed special-purpose digital computers for supplying the final navigational fix in printed form, essentially as soon as the pass of the satellite is complete.

In summary, a typical sequence of operations that might occur in an operational Doppler satellite navigation system is outlined as follows.

1) Tracking stations receive, reduce, and record in coded digital form, a vacuum Doppler curve for any pass of any satellite within receiving range (above the horizon). This information is transmitted to the computing center by, for example, commercial telegraph or TWX.

2) The computing center uses the vacuum Doppler data from many such passes for each satellite to calculate oribital data at 12-hour intervals for nominally one day in advance. This information is transmitted to the injection station.

3) Nominally twice a day, as each satellite passes within range of the injection station, the station transmits the orbital data and time correction to the satellite. These signals first erase the satellite memory unit and then read in new orbital parameters and a new time correction. Injection of data for a full day, at 12-hour intervals, provides the requisite accuracy of the orbital data and time correction, and provides for one failure to inject without incapacitating that satellite.

4) The satellite retransmits data received from the injection station once each minute, otherwise the two harmonically-related frequencies are transmitted continuously and unmodulated.

5) A navigator need only turn on his receiving equipment at a time when a satellite is due to pass within range. His equipment receives and records the vacuum Doppler data. From these data and with time and frequency corrections supplied by the time standard from the satellite, the navigator's latitude and longitude are computed.

The above section has attempted to outline very briefly the over-all concepts and functions of a satellite Doppler navigation system. In Section III, some of the details of the computing procedures necessary to obtain the latitude and longitude are presented when they are pertinent to the error analysis of the navigational fix. Other than to note that all equipment and hardware are well within the present state of the art, no details of the instrumentation are discussed in this report. Some aspects of the instrumentation are discussed in other papers appearing in this same issue.[7]

III. Proof of Feasibility and Error Analysis

When the satellite orbit is known, a measurement of the Doppler curve in the vicinity of zero Doppler shift allows a determination of the time and the line-of-sight range to the observer at the point of closest approach. Consequently, the position of the navigator could be determined relative to the satellite orbit and, hence, his position on the Earth. The fact that the Doppler shift caused by motion of the satellite transmitter contains this information is well known. This principle has been used for years to determine the miss distances of CW homing missiles, and satellite tracking as well as navigation proposals have been made utilizing this principle. What has not been generally recognized is that by a more generalized and sophisticated analysis,

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[5] W. H. Guier, "Numerical computing problems in the Doppler tracking of near earth satellites," Proc. Meeting Math. Committee USE Organization, San Francisco, Calif.; August 1, 1958.

[6] W. H. Guier and G. C. Weiffenbach, "The Doppler determination of orbits," Proc. Natl. Aeronaut. and Space Admin. Conference on Orbit and Space Trajectory Determination, U. S. Naval Research Laboratory, Washington, D. C.; March 12, 1959 (in press).

[7] G. C. Weiffenbach, "Measurement of the Doppler shift of radio transmissions from satellites," this issue, p. 750.

information contained in the whole of the Doppler shift can be extracted to yield navigational accuracies competitive with the best existing navigation systems.

Two factors contribute to the accuracy which can be achieved by satellite Doppler navigation. The first is the well-known fact that frequency and time are two of the physical variables most readily measured with very high precision. The second factor, and this is the one that appears to have been largely overlooked, is the fact that there is tremendous redundancy in the information contained in the shape of the entire Doppler curve. The position of the navigator can be determined not only from the time and slope of the curve at the point of maximum slope, but also from a comparison of the Doppler shift at any two points on the curve. Thus, in principle, each pair of data points on the entire Doppler curve can be considered to provide a determination of the navigator's position and each of these determinations is more or less statistically independent. In practice, the optimum use of this redundant data is accomplished through the employment of curve fitting techniques such as the method of least squares.

The above, somewhat idealized, picture is, of course, true only when the errors are statistical in nature. Accuracy can be gained by adding more and more data points only up to the point where unknown biases in the data produce errors comparable to the accuracy obtained by using redundant data. For this reason, considerable emphasis has been placed on error analyses which include errors of a nonstatistical nature such as frequency drift, ionospheric refraction, etc.

It is impossible within the scope and length of this paper to report in detail on all the error studies that have been made to date by the authors and their colleagues. Consequently, reporting of results has been limited to those studies which lend understanding to the present approach to satellite Doppler navigation, establish the theoretical feasibility of obtaining accuracies to better than one-half mile, and exhibit sources of error that place limitations on the quality of instrumentation that is necessary. In Section III-A, the probable error arising only from errors of a statistical nature is given as a function of the ground range. In Section III-B various possible errors of a nonstatistical nature are considered.

A. Errors Contributed by Random Noise—Ultimate Accuracy

If one assumes that all the factors which might contribute to systematic or biasing errors can be accounted for, the accuracy of the system is only limited by the random errors arising from experimental noise. Consequently, an error analysis was performed in which all sources of error were ignored except the fundamental experimental random noise—the underlying thought being that the accuracy achievable under these conditions should roughly indicate the ultimate accuracy of the system.

Since Doppler navigation is based upon a frequency measurement, experimental noise appears as random errors in the measured frequency as a function of time. Estimates indicate that for a satellite at a 400 mile altitude with transmitter powers of about 100 mw, the RMS frequency error for any given statistically independent data point should be less than 0.5 cps at 100 mc, or about two to three parts in 10^9. This estimate was made assuming that "white" frequency noise existed and that the circuit time constants were such that a statistically independent data point could be taken once every second. However, the analysis given below did not utilize the maximum number of data points allowable under this assumption because 1) the maximum number of data points is so large (between 600 and 800) that computation methods available at the time were not adequate to handle so much data in a reasonable computing time, 2) approximate scaling laws exist for estimating the error for any number of data points, once the computations are performed with sufficient data points to provide a valid sample of the noise, and 3) it was not known at the time of the analysis what the autocorrelation time of the frequency noise would be in practice, and to be conservative, only data points separated by more than ten seconds were considered statistically independent. The results of the error analysis made under the conditions stated above will now follow.

The equations for the Doppler shift, as a function of the navigator's position, transmitter frequency, and time, are given in the Appendix. If the navigator is moving relative to the Earth during the time of the satellite pass, his position and his navigation error must be carefully defined. The main emphasis in this report is placed upon surface navigation (ships at sea), and therefore, the navigator's velocity is low and should be known with reasonable accuracy. Consequently, the navigator's position is defined to be his geodetic latitude and longitude at some chosen time t_0 during the pass of the satellite. With this definition, the Doppler shift has been written in the Appendix in terms of the quantities

ϕ_0, λ_0 = geodetic latitude and longitude of the navigator at the time t_0,

$\delta\phi(t), \delta\lambda(t)$ = change in latitude and longitude of the navigator during the pass time of the satellite relative to ϕ_0 and λ_0.

It was assumed throughout, unless otherwise stated, that $\delta\phi(t)$ and $\delta\lambda(t)$ contained negligible error, and that the navigation error was defined to be the error resulting in ϕ_0 and λ_0.

The error analysis was divided into two parts. First, a theoretical study was made based upon the usual techniques of curve fitting analysis. In this analysis it was assumed that about 50 data points were used, and the RMS frequency error was chosen to be 5 parts in 10^9. The results of this analysis are indicated in Fig. 1 as the solid, dashed, and dash-dot curves.

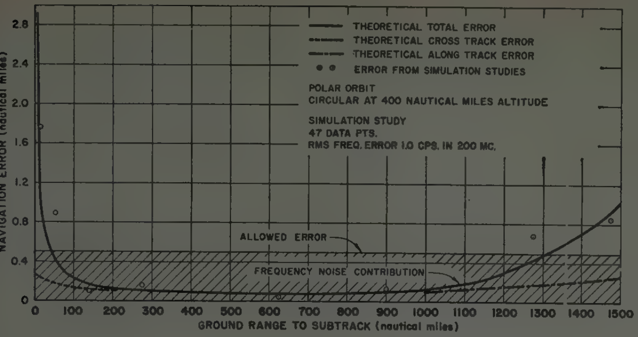

Fig. 1—Contribution of frequency noise to navigational error, as a function of ground range.

Second, a simulation was made in the following way. The expected Doppler frequency curve was computed assuming some position (ϕ_0 and λ_0) for the navigator. From this "theoretical" Doppler curve, 47 data points were chosen and random numbers, representing experimental noise, were added to the theoretical data to form synthetic experimental data. The magnitude of the noise was chosen to be the equivalent of a frequency error of about 5 parts in 10^9. Then, forgetting the position initially chosen for the navigator, his location was estimated to be that value of ϕ_0 and λ_0 which produced the least squares fit between the theoretical Doppler curve and the synthetic experimental data. In this way, the error in position introduced by the experimental noise could be estimated through a direct simulation of the navigator's computations and demonstrated that in practice computing procedures could be designed to achieve the surprisingly low error that is possible. The results of this simulation are indicated as circled dots in Fig. 1.

The results presented in Fig. 1 show the general characteristics that have been found in all error studies where only random noise is present. First, it can be seen that the error for most ground ranges is sufficiently low so that random noise cannot be a significant factor in achieving navigational accuracies of one-half mile. Second, in the few samples tried where synthetic data were used and actual navigation computations performed, the results agree quite well with the theoretically estimated error, lending confidence to the computer programs designed for navigation. Third, the navigational error becomes large at ground ranges such that the satellite appears above the horizon only for short times, and also at small ground ranges where the satellite passes nearly overhead. This is as expected since 1) when the satellite is always on the horizon, the determination of along-track error (latitude for a polar orbit) is difficult, and 2) for overhead passes the determination of the cross-track error (longitude for a polar satellite) is difficult since the slant range changes very little for changes in cross-track position when near the plane of the satellite orbit.

In an absolute sense, Fig. 1 does not represent the ultimate accuracy if all possible data points are used, transmitter power is increased to its practical limit, and more sophisticated curve fitting techniques are used. For order-of-magnitude calculations, a useful approximation is the following: if the RMS frequency error for each independent data point is σ_f, and the number of data points is N, the errors given in Fig. 1 can be scaled by $\sqrt{47/N}\cdot\sigma_f/0.5$. For example, assume the ultimate to be about one independent data point per second, and the RMS noise level to be 0.05 cps. At a 500-nautical mile ground range with a 400-mile satellite, the pass lasts for approximately 12 minutes or about 700 data points are received. Using, from Fig. 1, a navigational error of 0.1 nautical mile for 47 data points and 0.5-cps noise level, a true ultimate accuracy could be stated to be of the order of $0.1\sqrt{47/700}\cdot 0.05/0.5 = 0.003$ nautical mile$\cong$20 feet. Clearly, this number is only of academic interest because, without going to extreme complexity of instrumentation and computing techniques to eliminate the obvious errors due to nonstatistical biases, the statistical errors will contribute negligible error to the total navigation error. Applications such as those described by R. R. Newton in this issue justify the extreme care necessary to make the system as accurate as possible.

In summary, it can be seen that, for all practical purposes, statistical errors will not contribute a significant error to satellite Doppler navigation (except at the two extremes in ground range) when the system is economically designed to yield navigation accuracies of 0.5 nautical mile accuracy. Several possible sources of nonstatistical or biasing errors will now be considered.

B. *Errors Contributed by Nonstatistical Biases*

From the foregoing discussion, it can be seen that the inherent error in the system is well below one-half mile if careful accounting is made of all nonrandom sources of error. Several such sources of error are considered that are nonstatistical in nature.

1) *Ionospheric Refraction:* A major source of error in all satellite radio tracking schemes is ionospheric refraction. For purposes of studying its effect upon the Doppler shift, the ionosphere can be replaced by an equivalent index of refraction.[1,8] Since the Doppler shift of the satellite signal is basically the time rate of change of its electromagnetic path length, the Doppler shift is altered from what it would be in the absence of the ionosphere.

In the Appendix the equations for the Doppler shift in the presence of the ionosphere are given where the ionospheric contributions are stated in terms of a power series in the inverse of the transmitter frequency f.

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[8] J. A. Stratton, "Electromagnetic Theory," McGraw-Hill Book Co., Inc., New York, N. Y., p. 327; 1941.

Above 200 mc, the first term of the power series is usually all that is important. Using the mathematical expression for the first term that is given in the Appendix, the ionospheric contribution to the Doppler shift can be calculated and a typical result is shown in Fig. 2. The contribution shown in Fig. 2 was based upon the assumption that the electron density is only a function of altitude with a maximum density of 10^6 electrons per cubic centimeter at an altitude of 50 nautical miles.

The navigational error produced by ionospheric refraction can be understood qualitatively by noting that the maximum slope of the Doppler curve is a rough measure of the slant range, and that refraction has a direct effect upon this slope. Since the effect of refraction is to decrease the slope, the refraction error will be such as to place the navigator further from the satellite subtrack than would be true if the ionosphere were absent. To illustrate, a computation was performed similar to the calculations described in Section III-A. A synthetic Doppler curve was computed as before, but a refraction contribution was introduced in addition to the random errors simulating noise. The position of the navigator was then determined assuming that this refracted synthetic Doppler curve was a true experimental curve which contained no refraction. The results indicated that at a ground range of 500 nautical miles and a transmitter frequency of 200 mc, the navigational error was approximately 2 nautical miles. Furthermore, it was found that the best attainable fit of the refracted synthetic Doppler curve by the unrefracted theoretical Doppler curve was about 2 cps RMS, as opposed to a fit of about 0.2 cps RMS when the refraction contribution was not included.

Studies on the effects of refraction such as above indicated that the refraction contributions could not be ignored for transmitter frequencies as high as 500 mc. Furthermore, a cursory glance at the degree to which the electron distribution in the ionosphere could be predicted, so as to make an *a priori* correction for refraction, indicated that such predictions were not sufficiently reliable to reduce the refraction errors dependably below one-half mile.[9] Since it was felt that it was inadvisable to use satellite transmitter frequencies of the order of 1000 mc for the navigation system, studies were made to determine the degree to which the ionospheric effects could be eliminated experimentally by the use of two transmitter frequencies and thereby take advantage of the dispersive effect of the ionosphere.

From the Appendix, it can be seen that the Doppler shift in the presence of the ionosphere is of the form

$$\Delta f(f \mid t) = -\frac{f}{c}\dot{\rho}(t) + \frac{\alpha(t)}{f} + \frac{\beta(t)}{f^2} + \cdots. \qquad (1)$$

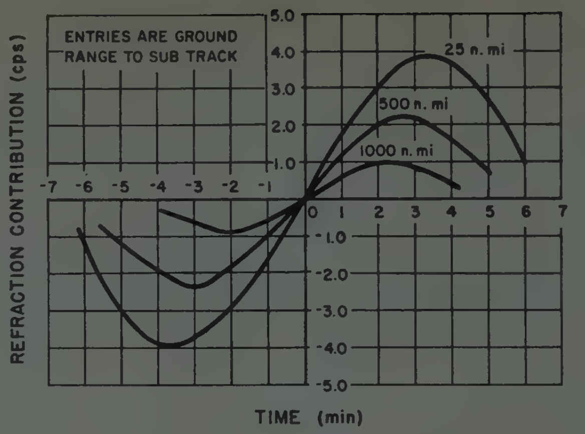

Fig. 2—Refraction contribution to Doppler shift, as a function of time, for several values of ground range. Transmitter frequency is 200 mc.

Noting that this form is true at each instant of time, it is clear that simultaneously measuring the Doppler shift at two different frequencies affords sufficient data to eliminate the first order refraction term, so that the result yields the vacuum Doppler shift plus only second order contributions of refraction. Eq. (2) indicates the trivial algebra necessary to eliminate the first order refraction contribution, and, of course, could be done in real time by frequency multipliers and a mixer provided f_1 and f_2 are harmonically related.

$$\Delta f_1(f_1 \mid t) = -\frac{f_1}{c}\dot{\rho} + \frac{\alpha}{f_1} + 0\left(\frac{1}{f_1^2}\right)$$

$$\Delta f_2(f_2 \mid t) = -\frac{f_2}{c}\dot{\rho} + \frac{\alpha}{f_2} + 0\left(\frac{1}{f_2^2}\right)$$

$$(f_2\Delta f_2 - f_1\Delta f_1) = -(f_2^2 - f_1^2)\frac{\dot{\rho}}{c} + 0\left(\frac{1}{f_{1,2}}\right). \qquad (2)$$

In these equations, $\dot{\rho}(t)$ is the time rate of change of the slant range from navigator to satellite as a function of time and contains all the information that is contained in the vacuum Doppler shift.

From the above discussion it can be seen that the effects due to refraction can contribute navigational errors only in second order when two frequencies are used, for example, 200 mc and 400 mc, and, consequently, a large step is made towards eliminating the ionosphere as a source of error. Table II gives an estimate of the relative magnitudes of refraction contributions as a function of frequency, and it can be seen that, based upon these data, two frequencies above 200 mc should be adequate to reduce the errors due to refraction to below 0.5 mile, remembering that the first order effect at 200 mc produced an error of about 2 miles.

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[9] E. L. Crow and D. H. Zacharisen, "Pilot Study of the Preparation of World Maps of F_2 Critical Frequencies and Maximum Usable Frequency Factors," National Bureau of Standards, Boulder, Colo.; NBS Rept. 5560; February 28, 1958.

In estimating the quantities appearing in Table II, no attempt was made to include effects such as winds and turbulence that can occur in the ionosphere, nor to include extremes that can occur at rare times during magnetic storms, etc. Consequently, there is still some question as to how small the residual refraction effects are under some circumstances. Nevertheless, it appears at the present time that for the large majority of cases the use of two frequencies can sufficiently eliminate the errors due to ionospheric refraction to below one-half mile.

TABLE II

REFRACTION CONTRIBUTIONS TO THE DOPPLER SHIFT

f (mc)	Vacuum term* (cps)	First order term* (cps)	Second order term* (cps)
50	1,250	10	0.5
100	2,500	5	0.08
200	5,000	2.5	0.01
500	12,500	1.0	—

* These are typical maxima of these quantities for a given Doppler curve and do not necessarily occur at the same time.

2) *Frequency and Frequency Drift Error:* Studies were made on the navigational error produced by instabilities in the frequency of the satellite transmitters and/or the navigator's local oscillator. First, it was found that when an unknown but constant frequency error occurred, no significant navigational error resulted so long as the navigator's position was computed by allowing the frequency of the satellite (more properly, the beat frequency between the satellite's transmitter and the navigator's local oscillator) to be an adjustable parameter, so that the navigational fix was determined by a three parameter fit to the experimental Doppler curve instead of a two parameter fit.

The fact that a three parameter fit produces no significant increase in navigational error over and above a two parameter fit, even when no frequency error exists, is due principally to the fact that the entire Doppler curve is used. When the entire curve is used, the shapes of the changes in the Doppler shift as a function of time that result from changes in frequency, latitude, and longitude, are completely different and, therefore, frequency, latitude, and longitude are "orthogonal" parameters. Consequently, the use of the whole of the Doppler curve not only provides redundant statistics for increasing the navigational accuracy, but also eliminates the error due to an unknown but constant frequency error. Because absolute accuracy of the various oscillators in the system would be expensive to achieve and maintain in practice, it was decided very early that a three parameter navigation calculation would always be necessary. Consequently, in all error analyses, including those quoted above, a three parameter calculation was made.

If either the satellite transmitters or the navigator's local oscillator change in frequency significantly during the pass of the satellite, the resulting navigation error is not so easily eliminated. Table III indicates the error that results from various drift rates of the oscillators, and again these figures were arrived at by simulating navigation (three parameter fit) using synthetic experimental data as outlined for previous calculations.

From Table III it can be seen that a stability of one part in 10^8 per hour is sufficient to allow navigation to one-half mile accuracy. Since such stabilities are not difficult to attain, no further attempt was made to study ways of reducing the errors by, for example, progressing to a four parameter fit.

TABLE III

NAVIGATION ERROR ARISING FROM FREQUENCY DRIFT
GROUND RANGE = 400 NAUTICAL MILES

Drift Rate (parts in 10^8/hr.)	Latitude Error (nautical miles)	Longitude Error (nautical miles)	Total Error (nautical miles)
0	0.08	0.1	0.13
1	0.12	0.19	0.23
10	0.15	0.80	0.86
50	1.13	2.86	3.08

3) *Error in the Navigator's Velocity:* An obvious source of error, which is of a nonstatistical nature, is an error in the navigator's velocity which, in effect, is an input to the computation of position. This error does not appear to be serious, however, unless special applications such as aircraft navigation are contemplated. It seems probable that a ship, for which accurate navigation to one-half mile is important, would know its velocity relative to the Earth to within one or two knots. Since the relative motion of the ship is only important during the pass time of the satellite (10 to 15 minutes), this means that the error in the position relative to ϕ_0 and λ_0 should be less than 0.5 nautical mile. Furthermore, as was the case for an unknown frequency, the alteration of the Doppler shift due to a velocity error of the navigator has a shape with time that is somewhat different than that caused by change of the parameters ϕ_0 and λ_0. Consequently, velocity errors produce changes in the Doppler curve that are partially orthogonal to the navigator's position. Therefore, the full effect of the velocity error is not felt in the navigational error. Typical navigational errors appear to be about one-half to two-thirds of the total error accrued in the ship's position during the pass time of the satellite, and, consequently, a two-knot velocity error produces about a one-fourth to one-third nautical mile navigational error.

Another possibility obviously exists for those ships for which accurate navigation is crucial. A crude autopilot could be used to measure velocity during the time that the satellite is above the horizon, and thereby reduce velocity errors to a negligible factor.

4) *Miscellaneous Sources of Error:* There are many sources of error that have not yet been mentioned. Such errors as uncertainty in the altitude of the navigator,

errors introduced by multipath reception of the signal (*e.g.*, reflection of the signal from the satellite by the water when the satellite is low on the horizon), the existence of more than one solution for the navigator's position, and many others, have been considered and do not appear to be serious. However, in studying them, several interesting facts have emerged which are worth mentioning.

First, altitude errors do not appear serious, since, for surface navigation, altitude (particularly at sea) is known to much better than a tenth of a mile. Such small errors make a negligible contribution. Signal propagation errors such as multiple path reception can be eliminated by building antennas that have nulls at the horizon, or by discarding data from the beginning and end of a pass. The amount of Doppler data lost by not receiving the signal below a few degrees elevation is not serious since, as emphasized previously, so many data points are received that the loss of these data are of little consequence.

Many sources of error that are significant to other types of navigation are not pertinent here. For example, since no angle measurements are involved, stabilizing of antennas is not required. Furthermore, no knowledge of the local vertical is needed, nor is a knowledge of true north.

Finally, an interesting point arises when considering the possibility that more than one position of the navigator yields the same fit of the experimental data. It was found that if the navigator was so ignorant of his approximate position that an attempt was made to find a position on the wrong side of the satellite subtrack, a minimum in the mean square fit of the Doppler data with respect to frequency, latitude, and longitude could be found. However, for this false minimum, the fit of the data was very poor as compared to the fit obtained on the correct side of the satellite subtrack. This poor fit arises because the rotation of the Earth gives the navigator an extra component of velocity relative to the satellite that changes sign when going from one side of the satellite subtrack to the other side. Consequently, if the navigator should find the false fit to the data, indicated by large residuals between the theoretical Doppler curve and the experimental data, the true position can be found by shifting to the other side of the satellite subtrack. To this date, no other false solutions have been found.

IV. Conclusions

It has been shown that when full use is made of the Doppler data, satellite Doppler navigation to an accuracy of one-half nautical mile is theoretically feasible. The necessary satellite instrumentation is attainable in the present state of the art. The satellites should contain two transmitters at different frequencies, say 200 and 400 mc, and should have a stability of about one part in 10^8 and power levels of about 100 mw. To supply the necessary inputs conveniently to the navigator and allow for updating the satellite orbits frequently, the satellites should contain a digital memory with means for receiving orbital data and time corrections from an injection station, and a means for serially reading out the memory and modulating at least one of the transmitters with these data together with a time synchronizing signal obtained from the satellite's oscillator. This allows the navigator to receive the time pulse with its correction and the orbital data necessary to perform the calculation of position. A system of three to four satellites in the proper orbits allows a navigation fix at least once every two hours.

It has been indicated that the main contributors to the navigational error are the various sources of biases, and not the fundamental experimental noise. Use of the whole Doppler curve, furthermore, allows for a three-parameter fit of the experimental vacuum Doppler data, the latitude and longitude of the navigator, and a vernier correction to the frequency. The use of a three-parameter fit makes it unnecessary to hold absolute accuracy of the satellite and navigator's oscillators. The second-order refraction effects that remain when two frequencies are used may be the major source of error, particularly in special circumstances when magnetic storms are present or when navigating in the auroral regions. However, for typical ionospheric conditions, the use of two frequencies to eliminate first order refraction effects is sufficient to allow one-half mile accuracy.

Navigation by such a satellite Doppler system has many obvious advantages. The system is all weather, passive, and accurate. The navigator does not need to have knowledge of local vertical or the direction of north. Since no angle data are needed, special antenna arrays are not needed nor do the problems of antenna bore-sighting or stabilization arise. Depending upon the navigational accuracy required, the navigator can employ various degrees of sophistication in his instrumentation and computing equipment.

In conclusion, one further remark is worth mentioning. Assuming the existence of a reliable world-wide navigation system which has an accuracy of one-half mile, its optimum use is mapped to at least this accuracy. As discussed in another paper in this issue, a more careful application of the same principles would provide the means for such accurate mapping of the world.

Appendix

Equations for the Doppler Shift in the Presence of the Earth's Ionosphere

This Appendix presents the derivation of the Doppler shift (including first-order refraction effects) as a function of the geodetic position of the navigator, the time, the satellite frequency, and the ionospheric electron density. The notation used is discussed, the major assumptions necessary for the derivation are listed, a derivation is given of the first two terms in the series expansion of the refracted Doppler shift in powers of one over the transmitter frequency, and finally, the detailed expressions for the vacuum Doppler shift are given.

A. Notation

t = time

$x_s(t), y_s(t), z_s(t)$ = Cartesian coordinates of the satellite in a right-handed system where the origin is at the earth's *CG*, the *x-z* plane coincides with the Greenwich meridian, and the *x-y* plane coincides with the earth's equatorial plane

$\phi(t)$ = geodetic latitude of the navigator

$\lambda(t)$ = geodetic longitude of the navigator

ϕ_0, λ_0 = geodetic latitude and longitude of the navigator at a time t_0, chosen to be some time during the pass of the satellite

a, f_0 = semimajor axis and flattening of navigator's datum

$x_n(t), y_n(t), z_n(t)$ = Cartesian coordinates of the navigator in the above coordinate system

$\rho(t)$ = instantaneous slant range between the navigator and satellite

$N(x, y, z)$ = electron density in the ionosphere per cubic centimeter

e = electronic charge in esu

m = electron mass in gm

$f_N(x, y, z)$ = ionospheric plasma frequency $= \sqrt{N(x, y, z)e^2/\pi m}$

$n(x, y, z|f)$ = equivalent refractive index of ionosphere

f = satellite transmitter frequency

c = velocity of light in vacuum

$P(t)$ = optical path of CW signal from satellite to navigator

$Pg(t)$ = geometric path of CW signal from satellite to navigator

ds = differential arc length along some defined path P

$0(\xi)$ = terms of order ξ or higher

$\dot{g}(t)$ = time derivative of $g(t)$

$\Delta f(f|t) = \Delta f(\lambda_0, \phi_0, f|t)$ = refracted Doppler shift.

B. Assumptions

1) The satellite transmitter frequencies are above 100 mc.
2) The effects of the ionosphere can be described to a sufficient approximation for $f \geq 100$ mc by ascribing to the ionosphere an equivalent index of refraction given by

$$n(x, y, z) = \sqrt{1 - \left[\frac{f_N(x, y, z)}{f}\right]^2},$$

and the phase velocity of the signals transmitted from the satellite is given by $v = c/n(x, y, z)$.

3) The change in altitude of the navigator during the pass of the satellite is negligible, *i.e.*, surface navigation is assumed.
4) Relativistic effects are negligible.

C. Refracted Doppler Shift Represented as a Power Series in 1/f

Using the above assumptions, when both transmitter and receiver are moving slowly compared to the speed of light, the Doppler shift is given by

$$\Delta f(f|t) = -\frac{f}{c}\frac{d}{dt}\int_{P(t)} n(x, y, z|f)ds(x, y, z), \tag{3}$$

where the optical path $P(t)$ is defined as that path for which

$$\int_{P(t)} n(x, y, z|f)ds(x, y, z)$$

is an extremum in comparison with all possible paths that begin at the satellite and end at the receiver. Consequently

$$\int_{P(t)+\delta P(t)} nds - \int_{P(t)} nds = 0(\delta P^2), \tag{4}$$

where $\delta P(t)$ is any path that lies close to $P(t)$.

To start the development of the Doppler shift, $\Delta f(f|t)$, in a power series in $1/f$, (4) can be applied by identifying $\delta P(t)$ with the small difference in the optical and geometric paths when $f \geq 100$ mc. Noting first that

$$\lim_{1/f\to 0} n(f) = 1 \quad \text{and} \quad \lim_{1/f\to 0} P(t) = Pg(t),$$

it is reasonable to suppose that $\delta P(t) = P(t) - Pg(t) = 0(1-n)$. Thus, when the refractive index n lies close to unity, as it does when $f \geq 100$ mc, (4) becomes

$$\int_{P(t)} n(x, y, z|f)ds(x, y, z)$$
$$= \int_{Pg(t)} n(x, y, z|f)ds(x, y, z) + 0[(1-n)^2].$$

Rewriting the above equation,

$$\int_{P(t)} n(x, y, z|f)ds(x, y, z)$$
$$= \int_{Pg(t)} ds(x, y, z) - \int_{Pg(t)} [1 - n(x, y, z|f)]ds(x, y, z)$$
$$+ [0(1-n)^2]$$
$$= \rho(t) - \int_{Pg(t)} (1-n)ds + 0[(1-n)^2], \tag{5}$$

where $\rho(t)$ is the geometric slant range from navigator to satellite, and is independent of $n(x, y, z|f)$.

Again, noting that for $f \geq 100$ mc,

$$\frac{f_N(x, y, z)}{f} \ll 1,$$

and

$$1 - n(x, y, z \mid f) = \frac{1}{2}\left(\frac{f_N}{f}\right)^2 + 0\left[\left(\frac{f_N}{f}\right)^4\right].$$

Thus,

$$\int_{P(t)} n(x, y, z \mid f) ds(x, y, z)$$

$$= \rho(t) - \frac{1}{2f^2}\int_{P_g(t)} f_N{}^2 ds(x, y, z) + 0\left[\left(\frac{f_N}{f}\right)^4\right]. \quad (6)$$

Substituting (6) into (3) yields the required power series to first order.[10]

$$\Delta f(f \mid t) = -\frac{f}{c}\dot{\rho}(t) + \frac{1}{2fc}\int_{P_g(t)} f_N{}^2(x, y, z) ds(x, y, z)$$

$$+ 0\left(\frac{f_N{}^4}{f^3}\right). \quad (7)$$

D. *Dependence of the Vacuum Doppler Shift on the Position of the Navigator*

It was established in Section III that it is sufficient to define the navigator's position to be the value of the geodetic latitude and longitude of the navigator at some instant of time t_0 that occurs approximately in the middle of the pass of the satellite.[11] Let this position be denoted by

$$\left.\begin{aligned} \phi_0 &= \phi(t_0), \\ \lambda_0 &= \lambda(t_0), \\ \text{and let} & \\ \delta\phi(t) &= \phi(t) - \phi(t_0), \qquad \dot{\phi}(t) = \delta\dot{\phi}(t), \\ \delta\lambda(t) &= \lambda(t) - \lambda(t_0), \qquad \dot{\lambda}(t) = \delta\dot{\lambda}(t). \end{aligned}\right\} \quad (8)$$

Then, under the above assumptions, to a sufficient approximation,

$$\left.\begin{aligned} \zeta_n(\phi_0) &= \frac{a}{\sqrt{\cos^2\phi_0 + (1 - f_D)^2 \sin^2\phi_0}} \\ \xi_n(\phi_0 \mid t) &= \zeta_n(\phi_0)\cos\phi(t) \\ \dot{\xi}_n(\phi_0 \mid t) &= -\left[\zeta_n(\phi_0)\sin\phi(t)\right]\dot{\phi}(t) \end{aligned}\right\} \quad (9)$$

$$\left.\begin{aligned} x_n(\phi_0, \lambda_0 \mid t) &= \xi_n(\phi_0 \mid t)\cos\lambda(t) \\ \dot{x}_n(\phi_0, \lambda_0 \mid t) &= \dot{\xi}_n(\phi_0 \mid t)\cos\lambda(t) - y_n(t)\dot{\lambda}(t) \\ y_n(\phi_0, \lambda_0 \mid t) &= \xi_n(\phi_0 \mid t)\sin\lambda(t) \\ \dot{y}_n(\phi_0, \lambda_0 \mid t) &= \dot{\xi}_n(\phi_0 \mid t)\sin\lambda(t) + x_n(t)\dot{\lambda}(t) \\ z_n(\phi_0 \mid t) &= \zeta_n(\phi_0)(1 - f_D)^2\sin\phi(t) \\ \dot{z}_n(\phi_0 \mid t) &= \xi_n(\phi_0 \mid t)(1 - f_D)^2\dot{\phi}(t). \end{aligned}\right\} \quad (10)$$

Using the above equations, the Doppler shift is given by

$$\begin{aligned} \rho(\phi_0, \lambda_0 \mid t) &= \left\{\left[x_s(t) - x_n(t)\right]^2 + \left[y_s(t) - y_n(t)\right]^2 + \left[z_s(t) - z_n(t)\right]^2\right\}^{1/2} \\ \dot{\rho}(\phi_0, \lambda_0 \mid t) &= \frac{(x_s - x_n)(\dot{x}_s - \dot{x}_n) + (y_s - y_n)(\dot{y}_s - \dot{y}_n) + (z_s - z_n)(\dot{z}_s - \dot{z}_n)}{\dot{\rho}(\phi_0, \lambda_0 \mid t)}. \end{aligned} \quad (11)$$

From (7) it can be seen that the zeroth order term is the vacuum Doppler shift and the first order term is a correction to the vacuum Doppler shift due to the presence of the ionosphere. Noting that the plasma resonance frequency f_N is proportional to the square root of the electron density, it can be seen that (7) exhibits the well-known fact that, to a good approximation, the refraction contribution to the Doppler shift is proportional to the time rate of change of the total number of electrons along the signal path.

Substituting (11) into the expression for the vacuum Doppler shift, $-f\dot{\rho}(\phi_0, \lambda_0|t)/c$, yields the required dependence of the Doppler shift on the geodetic position of the navigator.

ACKNOWLEDGMENT

The authors acknowledge the contributions of F. T. McClure, who originally suggested the application to navigation, and R. B. Kershner, who, with Mr. McClure, made most of the contributions to the concepts of the over-all system. R. R. Newton and many others at the Applied Physics Laboratory, Johns Hopkins University, also contributed to many stimulating discussions.

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[10] The above power series has been rigorously proved by one of the authors (W. H. Guier) to be correct to third order under the above assumptions and proved to first order when the second assumption has been removed. There appears no reason to believe that such a representation of the refracted Doppler shift to higher order is not valid.

[11] Clearly, for aircraft navigation, this definition must be generalized to be useful. While introducing no essential complications in theory, such considerations are beyond the scope of this paper.

Electrically-Controlled Demand-Thrust Rockets for Satellite Guidance*

DOUGLAS D. ORDAHL†

Summary—**Satellites which can be positioned precisely in space will be required for the space research programs of the near future. The information obtainable with such vehicles will be of immediate importance to both military and basic scientific research programs. Typical applications include meteorological observations, communications, atmospheric research and, ultimately, soft-landing vehicles for manned space travel. To accomplish accurate control of orbit and/or attitude with rocket motors requires extremely precise regulation of thrusts, generally in the range between 0 to 1 or 0 to 10 pounds. It must be possible to turn these control forces on and off in a few milliseconds repeatedly or intermittently over periods of from several months to perhaps several years. For use in such applications, rocket units have been investigated which employ the combination of storable-liquid propellants and demand-thrust injectors capable of complete remote control. An experimental model of a 0- to 10-pound thrust engine has been successfully demonstrated. Three types of electrically actuated control systems are being investigated. These systems employ direct solenoid control, a magnetic ram hydraulic booster, and a micropump with a hydraulic servomechanism. Guidance signals of 50 microamps are used to regulate power to the engine controls.**

INTRODUCTION

ATTAINMENT of the objectives of both the scientific and military space programs requires use of a rather large number of satellites and space vehicles which can be guided over relatively long periods of time after the main propulsion engines have separated or shut down. The applications include satellites for meteorological observations, communication relays, probes for atmospheric research, soft-landing vehicles for moon and planetary probes, and, ultimately, manned space travel. The types of guidance problems encountered include, 1) the relatively simple task of correcting elliptical orbits into the more stable circular paths, 2) the more sophisticated establishment of a preselected orbit, and 3) the very precise task of positioning a satellite with respect to orbital shape, direction, elevation, and attitude for observation of specific points on earth or areas in space.

Because of the well-publicized weightless, frictionless conditions which prevail, satellite or space vehicle guidance must be accomplished by the application of reaction forces including cases where such devices as gyro stablization may be used. In contrast to the extremely large thrust required to place a vehicle in orbit, the forces required for even relatively major corrections to vehicles weighing as much as 1000 pounds are only in the region of 10 pounds thrust. In addition, extremely precise control is required over the thrust level and action time to prevent errors of over-correction. Interpreting these problems in terms of the specific requirements that must be provided by the rocket system shows that 1) the response times in reaching thrust level must be no longer than 10 milliseconds, 2) the propulsive force must be delivered either in extremely reproducible short impulses or must respond rapidly and accurately to the demand for any required thrust level within the capacity of the engine, and 3) the system must be completely storable and capable of intermittent response for periods of as much as one year or more. In addition to these requirements the highest effective impulses available are required to maintain the weight and volume of the satellite vehicle at the practical minimum.

DEVELOPMENT OF DEMAND-THRUST ENGINES USING STORABLE LIQUID PROPELLANTS

In the past two years, research on variable-thrust storable-liquid propellant propulsion systems (undertaken originally in connection with advanced air-to-air missile work) resulted in the development of a simple engine which could be turned completely on or off or adjusted to any desired thrust level on demand. This development used storable hypergolic liquid propellants, pressure pumped to a variable area injector which maintains proper balance in the flow of an oxidizer and fuel at all thrust levels. Most of the work has been carried out using such propellants as unsymmetrical dimethylhydrazine and red fuming nitric acid; however, a rather wide variety of propellant combinations can be used successfully providing they are hypergolic with short ignition delays. Nonhypergolic propellant combinations could be used with the provision of an adequate ignition system. An example of the reliability and durability of the design is indicated by the typical experience with one 3500-pound thrust engine having an accumulated test time of over six hours with no signs of deterioration.

As the requirements for satellite and space vehicle guidance became known, the concept of the storable demand-thrust engine was extended to include a series of very small engines such as the 0- to 10-pound motor shown in Figs. 1 and 2. This particular engine weighs less than one pound and operates at a combustion chamber pressure of 200 psi. A series of test firings in the static test stand shown in Fig. 2 has demonstrated instantaneous ignition, smooth combustion, complete shut-off on demand without leakage, and no evidence

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† U. S. Naval Ordnance Test Station, China Lake, Calif.

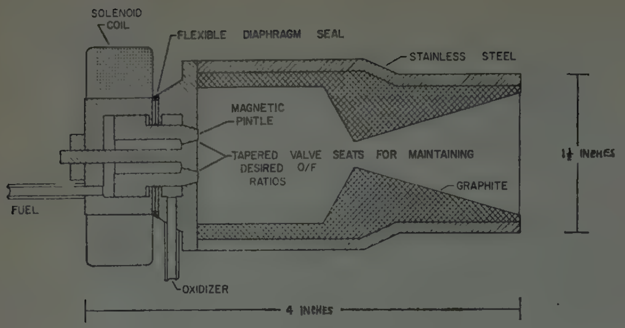


Fig. 1—Zero- to 10-pound variable thrust engine.

Fig. 2—Zero- to 10-pound variable thrust motor operating in the static test stand. Flame is colorless.

of deterioration after an extended period of operation. Response times with suitable electrical power are estimated to be less than 10 milliseconds. An interesting discovery is that in all the tests made the flame has been completely colorless except during start-up and shut-down. The motor shown in Fig. 2 is actually operating at full thrust.

Control Methods for Integrated Systems

An optimum design for a 0- to 10-pound demand-thrust engine is estimated to require no more than 0.1 watt of power for the electrical controls. However, the familiar problems of weight, size, and response optimization are encountered and proper selection of a control circuit depends to a significant degree on the particular application. Where suitable electrical power is available, direct solenoid actuation is a reasonable choice to achieve high rates of response. On the other hand, systems may be required which use hydraulic pressure or other forms of control to minimize power consumption.

With these considerations in mind, several other methods are under development for actuation of the demand-thrust motors. It should be noted at this point that some navigation systems will require linear thrust level control, particularly where closed-loop servos are employed, and some systems will require predetermined impulses to be delivered in a square wave pattern. In the system diagram shown in Fig. 3, pressurization of the liquid propellants is accomplished by a helium gas bottle, and, as the fluids are expelled, thin metal bags collapse to prevent sloshing, much as in a toothpaste tube. An amplifier and a secondary power supply are used to build up the control signal sufficiently to operate the solenoid coil at the desired rate of response. A typical power supply for this application would comprise a 12-volt nickel-cadmium battery weighing 0.75 pound and a solar cell recharge system which would weigh as much as 4 pounds. The magnetic amplifier would be of the 2- or 3-stage transistor type weighing 0.5 pound and furnishing 2 watts to the solenoid coil controlled by an input signal of 50 microamps. The solenoid coil used requires 1.5 watts for operation and weighs 0.25 pound.

The diagram shown in Fig. 4 illustrates the use of an electrohydraulic actuator for driving the injector. This design would use hydraulic pressure on an enlarged area on the base to move the injector pintle. The electrical power would drive the solenoid actuated piston to increase or decrease the hydraulic pressure as required. This pressure would oppose the fuel pressure acting to close the piston. Thus, as the position of the small solenoid or magnetic piston is retracted, the pressure drops and the pintle closes, reducing or stopping the thrust.

The system illustrated in Fig. 5 makes use of hydraulic control of the injector obtaining the hydraulic pressure from a micropump. Use of this system presumes that sufficient electrical power would be available from other components in the vehicle to operate the pump. The electrically actuated servo valve would be one of a type currently available and would require only 10 milliwatts from the guidance systems for control. The micropump used in this design requires a power input of 50 watts and produces a flow rate up to 0.025 gallon per minute at 1000 psi.

Typical Applications

Among the typical applications that have been studied for small demand-thrust motors, is the type of vehicle shown in Fig. 6. Such a vehicle will use 12 small

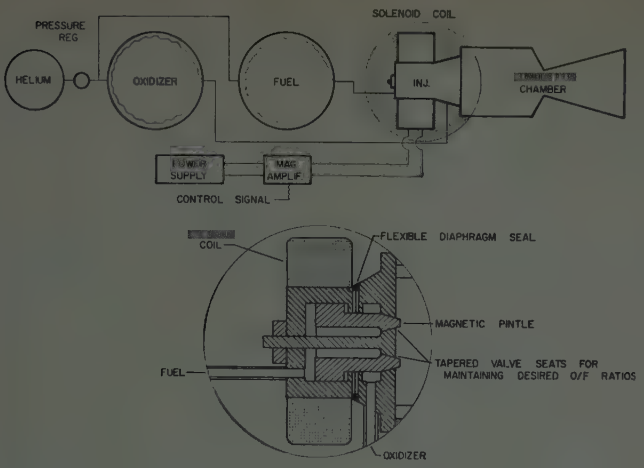

PRESSURE REG
HELIUM
OXIDIZER
FUEL
SOLENOID COIL
INJ.
CHAMBER
POWER SUPPLY
MAG AMPLIF
CONTROL SIGNAL
COIL
FLEXIBLE DIAPHRAGM SEAL
MAGNETIC PINTLE
TAPERED VALVE SEATS FOR MAINTAINING DESIRED O/F RATIOS
FUEL
OXIDIZER

Fig. 3—Zero- to 10-pound variable thrust propulsion system with direct electrical actuation.

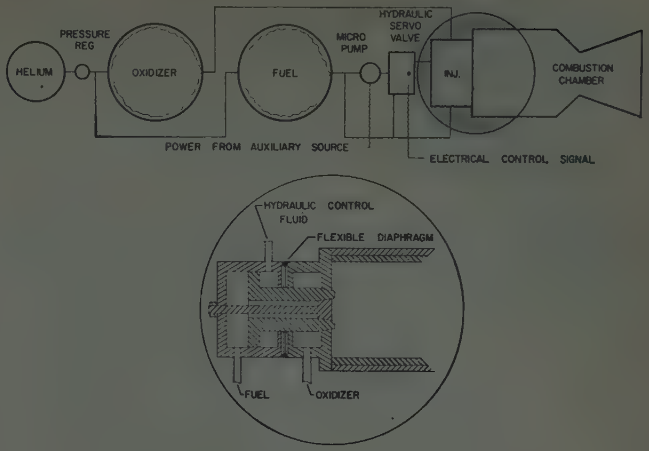
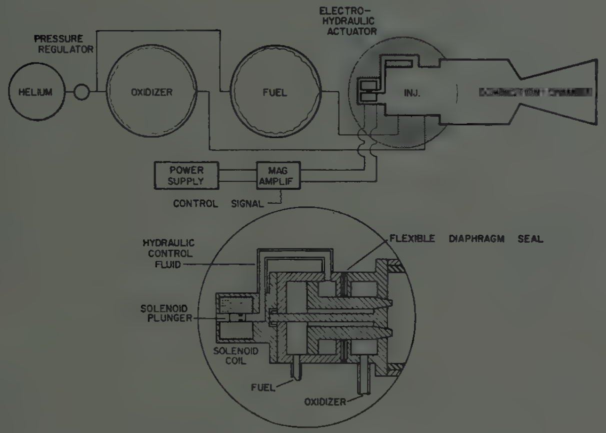

PRESSURE REG
HELIUM
OXIDIZER
FUEL
MICRO PUMP
HYDRAULIC SERVO VALVE
INJ.
COMBUSTION CHAMBER
POWER FROM AUXILIARY SOURCE
ELECTRICAL CONTROL SIGNAL
HYDRAULIC CONTROL FLUID
FLEXIBLE DIAPHRAGM
FUEL
OXIDIZER

Fig. 5—Zero- to 10-pound variable thrust propulsion system with hydraulic servo controls.


PRESSURE REGULATOR
HELIUM
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MAG AMPLIF
CONTROL SIGNAL
HYDRAULIC CONTROL FLUID
FLEXIBLE DIAPHRAGM SEAL
SOLENOID PLUNGER
SOLENOID COIL
FUEL
OXIDIZER

Fig. 4—Zero- to 10-pound variable thrust propulsion system with electrohydraulic actuation.

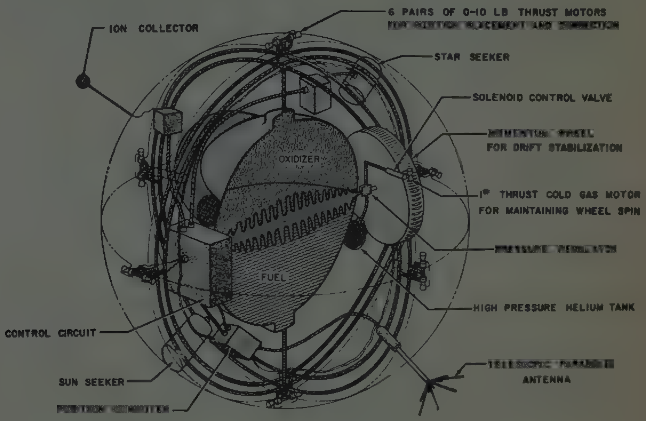
ION COLLECTOR
6 PAIRS OF 0-10 LB THRUST MOTORS
STAR SEEKER
SOLENOID CONTROL VALVE
FOR DRIFT STABILIZATION
OXIDIZER
1# THRUST COLD GAS MOTOR FOR MAINTAINING WHEEL SPIN
FUEL
HIGH PRESSURE HELIUM TANK
CONTROL CIRCUIT
SUN SEEKER
ANTENNA

Fig. 6—Typical research satellite using six pairs of opposed demand thrust jet motors for position control.

engines for position or short term attitude control. A typical application is for a 22,300-mile or 24-hour orbit civil communications satellite to maintain a fixed position. To permit the use of a parabolic antenna for minimizing transmitter output, such a satellite would need to be at least approximately oriented toward appropriate ground receivers. Other possible applications for vehicles such as the one illustrated are for earth and space scanning devices and scientific satellites which can be programmed into several different orbits.

Somewhat larger electrically controlled demand thrust engines (in the range of 5000 to 10,000 pounds thrust) would satisfy the requirements for a soft-landing vehicle to place an instrumented payload on the moon and would ultimately be useful in bringing one or more satellites together at a point in space. Such vehicles which can maneuver effectively in space would require a main lift (or retarding) motor and several smaller units for attitude control. Demonstration vehicles have been designed which could be electrically or radio controlled from the ground in flights as high as 50,000 feet from the earth. These vehicles could be used primarily to carry out experimental investigations on the integration of control, actuation, and propulsion systems. The variable-thrust propulsion units required for this type of experimental vehicle have already been designed and tested. In addition, the operation of injectors for engines of 0- to 20,000-thrust capacity have been demonstrated successfully.

A final application of immediate interest to the missile and space vehicle designer is that the small demand thrust motor shown in Fig. 1 uses so little propellant that it is suitable for a number of detailed high altitude studies of motor and propellant performance using relatively inexpensive vacuum chambers to get much of the data which was heretofore possible only from large elaborate installations. This unit is being used to investigate advanced designs for electrically actuated control systems and propellant performance characteristics at simulated altitudes up to 250,000 feet.

Inertial-Guidance Limitations Imposed by Fluctuation Phenomena in Gyroscopes*

GEORGE C. NEWTON, JR.†, MEMBER, IRE

Summary—**Because of recent improvements, inertial guidance is becoming increasingly important to space and missile programs in connection with the navigation of mobile bases and the launching of vehicles. Gyroscopes are an important component of inertial systems. In order to see if much room remains for further improvement, the performance limitation imposed by thermal fluctuations in gyroscopes is examined. This is done because experience in electronics shows that thermal noise is the ultimate performance limitation in sensitive instruments such as radio receivers and galvanometers. It is found that a system limited only by thermal noise in its gyroscopes would have several orders of magnitude more precision than that indicated in published performance data. This must mean that other factors are limiting performance of gyroscopes, and investigation of several of these shows that bearing noise may be an important one. In conclusion, since gyroscopes do not yet come near to attaining the performance limits set by thermal noise, there may be much room left for their further improvement.**

INTRODUCTION

DURING the past decade, the concept of inertial guidance has been transformed from a vision in the minds of a few pioneering engineers and scientists to a practical reality that must be taken into account in the planning of all present and future space programs. The initial use of inertial guidance has been largely restricted to earth-bound vehicles in the form of aircraft, submarines, and, most recently, ballistic missiles. How large a role inertial systems of navigation will play in the space programs of the future, and the exact functions that may be performed by such systems, will be determined, in part, by the accuracies that can be obtained. Therefore, it is important to know how accurate such systems are likely to become as the result of further improvements in their basic components. This paper analyzes the performance limitations imposed on inertial-guidance systems through fluctuation phenomena in gyroscopes and their associated stable platforms—two basic and related elements in such systems.

An intertial guidance system is used to determine continuously the velocity and position of a vehicle relative to some external frame of reference but without the aid of information from external sources. Basically, this may be done by carrying in the vehicle a local reference frame that bears a known relationship to the external one. Vehicle accelerations measured in this local reference frame can be processed to obtain velocity and position information referred to the external coordinates. For example, if the two frames were sets of rectangular axes maintained parallel to each other and the external one was fixed in inertial space, then the velocity and position of the vehicle could be obtained by successive integrations of accelerations measured along the axes carried by the vehicle. However, because instruments for measuring acceleration are also sensitive to gravity, it is necessary to compensate for this factor. Furthermore, it is frequently desirable to navigate in terms of an external-coordinate system moving relative to inertial space. This is expecially true of earth-bound vehicles such as aircraft, submarines and missiles in which navigation is done relative to earth coordinates. Gravity and the need to navigate in moving coordinates are two factors which greatly complicate inertial navigation systems.

In order to provide an internal-coordinate system that bears a known relationship to the navigational (or external) coordinates, it is generally convenient to use some form of stable platform inside the vehicle. A stable platform isolates the internal reference frame from angular motion of the vehicle. The platform itself may act as a local reference frame and have accelerometers mounted directly on it, or it may be simply a stabilizer for a reference frame that exists numerically in a computer. In certain systems for aircraft and submarine navigation, the stable platform may be arranged to seek the local vertical; then, in effect, by comparing this vertical with the initial vertical, the angular displacement of the vehicle along its desired track around the earth can be determined. Motion perpendicular to this track can be determined from integration operations applied to a cross-track accelerometer signal.

The above is a brief description of the inertial-guidance concept. Because of the many possible configurations that guidance systems may take and because of the complexities of any practical system, this brief description will have to suffice for purposes of this paper. The interested reader may refer to the literature[1–3] for more details. The extent to which inertial-guidance concepts will be used in space flight is critically dependent upon the accuracy that can be achieved. With present accuracies, it is doubtful if inertial schemes could be used exclusively for the navigation of space-

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* Original manuscript received by the IRE, December 2, 1959.

† Electronic Systems Lab., Dept. of Elec. Engrg., Massachusetts Institute of Technology, Cambridge, Mass.

[1] C. F. O'Donnell, "Inertial navigation," *J. Franklin Inst.*, vol. 266, pp. 257–277, 373–401; October and November, 1958

[2] J. M. Slater and D. B. Duncan, "Inertial navigation," *Aeronautical Engrg. Rev.*, vol. 15, pp. 49–52, 57; January, 1956.

[3] W. Wrigley, R. B. Woodbury, and J. Hovorka, "Inertial Guidance," Institute of Aeronautical Sciences, SMF Fund Paper No. FF-16 presented in January, 1957.

craft from one orbit to another. However, elements of such systems, *e.g.*, the stable platform, may be useful for isolating star-tracking devices from vehicle motion and for the other functions analogous to the autopilot of a conventional aircraft. For manned spacecraft it may be necessary to provide artifical gravity through vehicle rotation. Such rotation would necessitate a more elaborate type of autopilot than ordinary aircraft use, and some form of stable platform would be most useful.

PRECISION OF STABLE PLATFORMS

The heart of an inertial-guidance system is the stable platform which establishes the coordinate system for making acceleration measurements. Errors in the position of the stable platform would be directly reflected into errors in the computed position if the platform failed to maintain its prescribed position in relation to inertial space or the navigational coordinate system. Because of the key importance of the stable platform as a subsystem in an inertial navigator, an autopilot or a star-tracker, the remainder of this paper will be devoted exclusively to the question of platform positional accuracy.

The accuracy of a stable-platform system is largely determined by its *precision*. This situation results from the fact that great pains are taken to eliminate systematic errors through calibration procedures and through elaborate error compensation schemes. As the result of the care taken to eliminate systematic errors and as the result of the great improvement that has been made in the precision of the gyroscopes used to sense platform angular deviations, it has been possible to achieve accuracies in platforms that are satisfactory for many of the purposes for which they may be used in earth-bound vehicles.

Consider one of the angular coordinates of a stable platform θ_p. Fig. 1 shows the characteristic behavior of an ensemble of platforms 1 through n. Although initially aligned so that the angles are zero at $t=0$, at some later time T_r, there is a distribution in platform positions. Assuming no systematic errors, the average position across the ensemble will, of course, be zero. However, any particular platform will deviate from the average position of zero in an unpredictable manner. The standard deviation σ_{θ_p} will be used in this paper to characterize the spread of platform positions about the mean value. This standard deviation will be found to contain an unbounded term which builds up as the square root of time. Attention will be focused on this term to the exclusion of all others, since for large values of T_r it will predominate. In practice, when the standard deviation of the platform position has grown to a maximum permissible level, it will be necessary to reset the platform back to zero by means of auxiliary equipment such as a star sight.

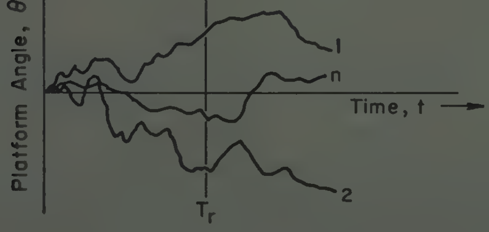

Fig. 1—Random behavior of an ensemble of stable platforms.

OPERATION OF STABLE PLATFORMS

A stable platform is basically a carriage for accelerometers or other equipment which is isolated from the vehicle's angular motions by means of appropriate gimbals. At least two, and sometimes three, gimbals in addition to the platform itself are used to permit the vehicle to have sufficient freedom for angular maneuver without disturbing the platform. If the platform suspension system were completely free of all friction and if the gimbals making up the suspension system were without inertia, then the mass of the platform would maintain its initial alignment in inertial space irrespective of vehicle maneuvers. However, in a practical system, three or four servomechanisms are used to drive the gimbals and platform in order to overcome friction and inertia effects. These servomechanisms are operated by error signals derived from gyroscopic instruments mounted on the platform.

In order to study the fundamental behavior of a stable platform, a simplified version will be used. Fig. 2 shows a single-degree-of-freedom platform driven by a single servomotor and amplifier from an error signal derived from a single-degree-of-freedom gyroscope of the integrating type. The function of this platform is to remain stationary in space in spite of vehicle maneuvers calling for angular rotation around the vertical axis c_3. It will be assumed that there is no angular motion around either of the other two orthogonal axes c_1 and c_2. Fig. 3 is a more detailed view of the integrating gyroscope which is used to sense angular motion of the platform and thereby to provide an electrical signal for actuating the platform servomechanism. This gyroscope operates on the principle that a torque about the gimbal axis (p_2 in Fig. 2) is required to balance the precession of the angular momentum of the gyroscope caused by a platform angular rate. This balancing torque is provided by viscous friction between the gimbal and the instrument case, which is rigidly attached to the platform. This means that an angular rate of the platform will cause an angular rate of the gimbal. However, any tilt angle θ_g of the gimbal which is built up as the result of the gimbal angular rate is detected by an electrical pickup which sends a signal to the amplifier driving the servomotor. As a result of this signal, the servomotor causes the platform to "unwind" and thereby restore the gimbal to its neutral position. By design, the tilt angle is kept small for all normal operating conditions.

Other kinds of gyroscopes could be used for sensing platform angular motion and generating correction signals. However, the single-degree-of-freedom integrating gyroscope shown in Figs. 2 and 3 is commonly used for this purpose in today's art. The perfection of this type of instrument has brought about a considerable improvement in the performance of stable platforms. This improvement has resulted primarily from the flotation of the gimbal by the damping fluid, thereby making possible a nearly frictionless suspension of the gimbal in the instrument housing. That is, flotation has eliminated to a very large degree coulomb-friction effects which previously introduced large uncertainties in instrument and platform performance. It is not the purpose of this paper to go into the intricacies of the analysis and the design of integrating gyroscopes, these being well covered by the literature.[4]

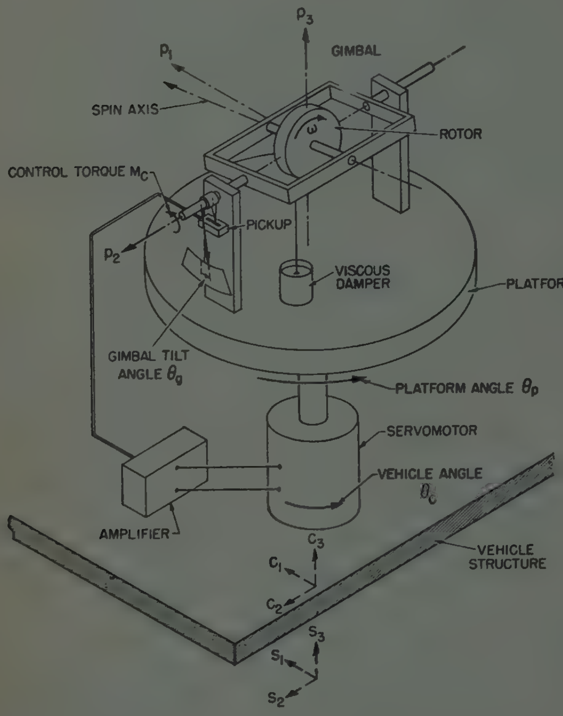
P3
P1
GIMBAL
SPIN AXIS
ROTOR
ω
CONTROL TORQUE M_C
PICKUP
P2
VISCOUS DAMPER
PLATFORM
GIMBAL TILT ANGLE θ_g
PLATFORM ANGLE θ_p
SERVOMOTOR
VEHICLE ANGLE θ_c
AMPLIFIER
C3
C1
C2
VEHICLE STRUCTURE
S3
S1
S2

Fig. 2—Schematic diagram of gyroscopically-stabilized single-degree-of-freedom platform.

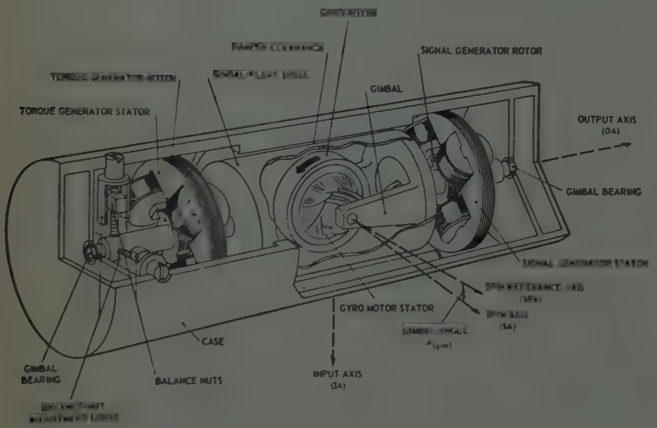
GYRO ROTOR
DAMPER CLEARANCE
SIGNAL GENERATOR ROTOR
TORQUE GENERATOR ROTOR
GIMBAL FLOAT SHELL
GIMBAL
TORQUE GENERATOR STATOR
OUTPUT AXIS (OA)
GIMBAL BEARING
SIGNAL GENERATOR STATOR
SPIN REFERENCE AXIS (SRA)
GYRO MOTOR STATOR
SPIN AXIS (SA)
GIMBAL ANGLE $A_{(gim)}$
CASE
GIMBAL BEARING
BALANCE NUTS
INPUT AXIS (IA)
BALANCE NUT ADJUSTMENT FORKS

Fig. 3—Pictorial view of floated, integrating gyroscope. (Courtesy, M.I.T. Instrumentation Laboratory.)

In order to clearly define the angles which will be referred to throughout the remainder of the paper, Fig. 4 shows three sets of rectangular coordinate axes. The set s_j refers to axes fixed in inertial space. The set c_j is attached to the vehicle and the set p_j to the platform. Because we are considering a simplified case of a single-degree-of-freedom platform, the axes s_3, c_3, p_3 are assumed aligned. The angle θ_p is the angle between the s_2 axis and the p_2 axis; this angle measures how far the platform has turned with respect to inertial space. The angle θ_c measures how far the vehicle has turned with respect to inertial space and is the angle between the s_2 and c_2 axes.

Analysis of Platform Behavior

Ideally, the system shown in Fig. 2 should operate to produce a platform angle at all times equal to the desired angle which in this paper will be assumed to be zero. In practice, the platform angle θ_p will depart from zero with the passage of time. This departure is random, since all systematic errors are assumed to be eliminated, and it is most easily characterized in terms of the standard deviation σ_{θ_p} of an ensemble of platforms.

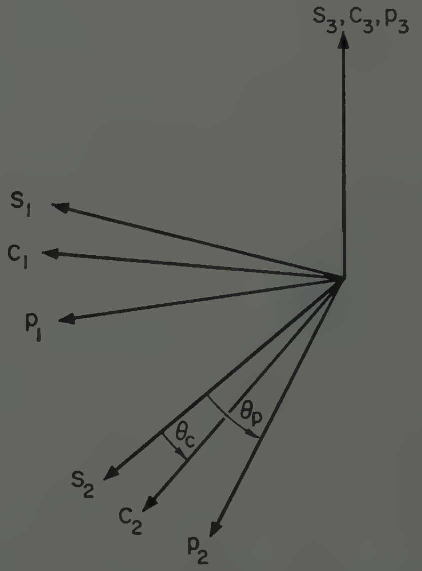
S_3, C_3, P_3
S_1
C_1
P_1
θ_p
θ_c
S_2
C_2
P_2

Fig. 4—Orientation of reference axes showing platform angle θ_p and vehicle angle θ_c.

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[4] C. S. Draper, W. Wrigley, and L. R. Grohe, "The Floating Integrating Gyro and its Application to Geometrical Stabilization Problems on Moving Bases," Institute of Aeronautical Sciences, SMF Fund Paper No. FF-13; presented in January, 1955.

How do factors such as thermal fluctuations in the gyroscope, spin-axis bearing noise, servomotor-torque fluctuations, and vehicle angular maneuvers affect the standard deviation σ_{θ_p}? The analysis that answers this question is based on Fig. 5, which shows a block diagram of the platform system of Fig. 2. In this diagram, s stands for the Fourier-transform complex variable. The summing point within the integrating gyroscope block represents torque summation around the gimbal axis. A control torque M_c for positioning the platform is balanced against the gyroscopic torque to yield a net torque which tilts the gimbal. However, in order to account for thermal and other fluctuations, a torque fluctuation M_f is added at this point. The net torque tilts the gimbal at a rate determined by the viscous-damping coefficient f_g. This tilt velocity occurs with a time lag τ_g which is equal to the moment of inertia J_g of the gimbal about its axis divided by f_g. Integration (represented by division by s) of the tilt rate yields the gimbal tilt angle θ_g.

The servomechanism of Figs. 2 and 5 is assumed to be a dc motor which is directly connected to the platform and whose armature voltage is controlled by an amplifier. Stabilization is provided by the armature back emf or by its synthetic equivalent derived from tachometric measurement of the armature velocity. The motor-driving torque is assumed to be linearly related to the gimbal tilt angle and is given by $k_g\theta_g$, where k_g is the over-all gain of the pickup and amplifier of Fig. 2. This torque is summed with a platform torque disturbance and a torque equivalent to the vehicle's rate of change of angle in order to form the net torque available to drive the platform relative to inertial space. The platform-torque disturbance accounts for noise in the servoamplifier, friction effects and load disturbances acting on the platform. The net torque divided by the effective damping coefficient f_p of the motor gives the platform's angular velocity Ω_p, providing the time lag associated with the motor and platform moment of inertia J_p is accounted for. The time constant τ_p is the inertia J_p divided by the damping f_p. It is assumed that the reaction torque of the motor on the vehicle produces negligible motion of the latter. The angular rate Ω_p of the platform feeds back to the gyroscope gimbal through the angular momentum H of the gyro rotor. This closes the platform-stabilization loop. The platform angle θ_p is simply the integral of the angular rate Ω_p.

Long-term errors in the platform angle can be corrected for by a reset mechanism, schematically shown in Fig. 5. Assuming that some independent means of determining a space reference angle θ_s, such as a star-tracker, is available, the error between this angle and the platform angle can be used to generate a control torque M_c which drives the platform (through the gyroscope) to the null condition. For reasons of convenience or necessity, the external reference angle may be available only intermittently, and this accounts for the switch in Fig. 5. Another way of asking the question concerning fluctuation effects is: How long may the period T_r between platform resets be before the standard deviation σ_{θ_p} of the platform angle exceeds an assigned limit?

The block diagram of Fig. 5 can be manipulated into the form of Fig. 6, where the influences of the several sources of platform-rate fluctuations are indicated by means of transmission functions terminating at a common summing point to form the platform rate Ω_p. From Fig. 5, these transmission functions are found to be

$$\begin{aligned} W_g(s) &= \frac{H^{-1}}{K_v^{-1}\tau_g\tau_p s^3 + K_v^{-1}(\tau_g+\tau_p)s^2 + K_v^{-1}s + 1} \\ W_d(s) &= \frac{(K_v f_p)^{-1}(\tau_g s + 1)s}{K_v^{-1}\tau_g\tau_p s^3 + K_v^{-1}(\tau_g+\tau_p)s^2 + K_v^{-1}s + 1} \qquad (1) \\ W_c(s) &= \frac{K_v^{-1}(\tau_g s + 1)s^2}{K_v^{-1}\tau_g\tau_p s^3 + K_v^{-1}(\tau_g+\tau_p)s^3 + K_v^{-1}s + 1} \end{aligned}$$

where the velocity constant K_v of the servomechanism is

$$K_v = \frac{k_p}{f_p} \qquad (2)$$

and the other constants are as defined above. (A summary of the nomenclature appears as Appendix I.)

Because it has been assumed that no systematic errors exist and that the desired platform angle is zero, only random fluctuations M_f of gimbal torque, platform torque M_d, and vehicle angle θ_c will be present. These fluctuations cause a random fluctuation of the platform angular rate Ω_p. Although these fluctuations have stationary probability distributions, those of the platform angle θ_p do not because initially the angle is

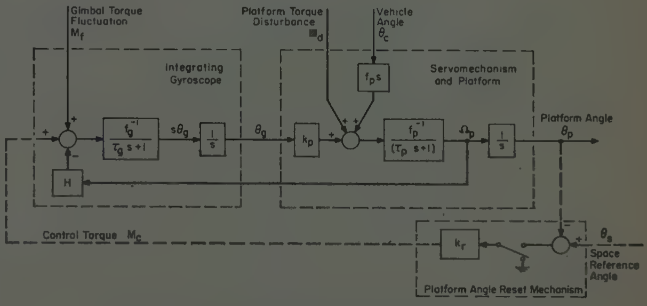
Gimbal Torque Fluctuation M_f
Platform Torque Disturbance M_d
Vehicle Angle θ_c
Integrating Gyroscope
Servomechanism and Platform
$f_p s$
+
+
$\frac{f_g^{-1}}{\tau_g s+1}$
$s\theta_g$
$\frac{1}{s}$
θ_g
k_p
+
+
+
$\frac{f_p^{-1}}{(\tau_p s+1)}$
Ω_p
$\frac{1}{s}$
Platform Angle θ_p
–
H
Control Torque M_c
k_r
–
+
θ_s
Space Reference Angle
Platform Angle Reset Mechanism

Fig. 5—Block diagram of stable-platform system.

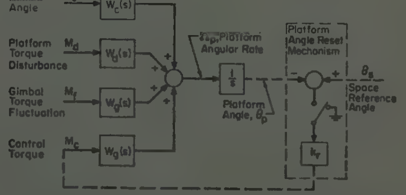
Angle
θ_c
$W_c(s)$
Platform Torque Disturbance
M_d
$W_d(s)$
Gimbal Torque Fluctuation
M_f
$W_g(s)$
Control Torque
M_c
$W_g(s)$
+
+
+
+
Ω_p, Platform Angular Rate
$\frac{1}{s}$
Platform Angle, θ_p
Platform Angle Reset Mechanism
–
+
θ_s
Space Reference Angle
k_r

Fig. 6—Simplified block diagram of stable-platform system.

set to zero and then, with the lapse of time, the angle wanders off from zero, as shown in Fig. 1. This situation is a nonstationary one and, unless a certain condition prevails, the standard deviation of the platform angle will grow without bound as a function of time because of the integration in going from rate to angle. This phenomenon is closely related to the random walk problem.[5]

The basic problem is best described in terms of Fig. 7 where a linear system characterized by its weighting function $w(t)$ is initially at rest. After $t=0$, it is connected to random input signal $x(t)$, which is drawn from a stationary process. $x(t)$ has zero mean value. The output $y(t)$, which was initially zero, assumes random values for time positive. How can the output be characterized as a function of time? By means of ensemble averages the rms value (or standard deviation) of the output can be expressed in terms of the auto-correlation function of the input and the system weighting function. In the frequency domain the result can be expressed as[6]

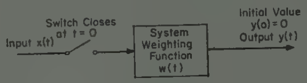

Fig. 7—Basic time-varying system used in platform stabilization.

$$\sigma^2(T_r) = \frac{1}{j}\int_{-j\infty}^{j\infty} ds W(-s, T_r) W(s, T_r) \Phi_{xx}(s) \qquad (3)$$

where $\Phi_{ss}(s)$ is the power-density spectrum of the input $x(t)$ and

$$W(s, T_r) = \int_0^{T_r} dt e^{-st} w(t) \qquad (4)$$

$W(s, T_r)$ is a time-varying transmission function.

A general analysis usually stops at this point because of the complexity of the time-varying transmission functions and the resulting difficulty in carrying out the integration (3). However, because the $w(t)$ of interest in the case of stable platforms is a simple integration, it is possible to carry the analysis further. As shown in Appendix II, the unbounded term in $\sigma^2(T_r)$ can be evaluated by contour-integration methods. Since all other terms are bounded, one obtains the general result

$$\lim \sigma(T_r) = [2\pi\Phi_{xx}(0)]^{1/2} T_r^{1/2} \text{ as } T_r \rightarrow \infty \qquad (5)$$

which shows that the long-term growth of the standard deviation of the output of an integrator is proportional to the square root of the product of time and the amplitude of the input power-density spectrum at zero frequency. This result will now be used to determine the standard deviation σ_{θ_p} of the platform angle for thermal noise in the gyroscope as the only fluctuation.

THERMAL NOISE

Just as a resistor is a source of thermal noise, so is a viscous damper in a mechanical system. The noise in systems containing resistors, whether mechanical or electrical, may be calculated if we assume each resistor to have a white-noise source embedded within itself. The power-density spectrum of the noise associated with a mechanical resistor or damper is

$$\Phi_{MfT}(s) = \frac{KT}{\pi} f_g. \qquad (6)$$

Here the subscripts stand for the thermal fluctuation moment, since this is the spectrum of the moment that thermal agitation produces within the damping fluid and that is effective in driving the gimbal. Here K is Boltzmann's constant and T is the absolute temperature; f_g is the coefficient of viscous damping associated with gimbal rotation.

If this thermal noise is the sole source of fluctuations, then from Fig. 6 we see that the power-density spectrum of the platform angular rate caused by thermal fluctuation moments is

$$\Phi_{\Omega pT}(s) = W_g(-s) W_g(s) \Phi_{MfT}(s). \qquad (7)$$

This relationship follows from the general properties of linear systems that are subjected to stochastic signals. Using (1), (6), and (7), one obtains

$$\Phi_{\Omega pT}(0) = \frac{KT}{\pi} \frac{f_g}{H^2} \cdot$$

Thus, from (5), the thermal noise influence on the platform position is expressed by

$$\sigma_{\theta pT}/\sqrt{T_r} = H^{-1}\sqrt{2KTf_g} \qquad (8)$$

providing that T_r is large. In these expressions, H is the angular momentum of the gyroscope rotor and the subscript T is used to denote thermal fluctuation.

Let us now establish an order of magnitude for the platform fluctuation using the above general result. Substituting in the parameter values for a typical gyroscope from Table I and putting in the numerical value for Boltzmann's constant, we get the following expression:

$$\frac{\sigma_{\theta pT}}{\sqrt{T_r}} = \frac{\sqrt{2(1.38 \times 10^{-16})350(2.0 \times 10^4)}}{1.0 \times 10^5}$$

$$= 0.440 \times 10^{-9} \frac{\text{rad}}{\text{sec}^{1/2}} \cdot \qquad (9)$$

This result can be made more meaningful by consideration of how long a time it would take for the standard deviation of the platform angle to grow to one minute of arc assuming that the only source of fluctuations is thermal agitation of the damping fluid. This time turns

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[5] S. Chandrasekhar, "Stochastic problems in physics and astronomy," *Revs. Mod. Phys.*, vol. 15, pp. 2–89; January, 1943.

[6] See, for example, J. H. Laning, Jr. and R. H. Battin, "Random Processes in Automatic Control," McGraw-Hill Book Co., Inc., New York, N. Y. pp. 238–239; 1956.

TABLE I

PARAMETERS FOR A TYPICAL INTEGRATING GYROSCOPE[7]

Rotor moments of inertia (around spin axis)	J_r	40 gm cm^2
Rotor mass	m_r	25 gm
Rotar angular velocity	ω_r	2500 rad/sec
Rotor angular momentum	H	1.0×10^5 dyne cm sec/rad
$\left[\text{Amplitude of power-density spectrum of bearing axial wander at zero frequency}\right]^{\frac{1}{2}}$	η_b	1.0×10^{-6} cm/(rad/sec)$^{1/2}$
Moment of inertia of gimbal (including enclosed equipment)	J_g	100 gm cm^2
Viscous-damping coefficient of gimbal	f_g	2.0×10^4 dyne cm sec/rad
Operating temperature	T	350° K

out to be 13,900 years. Thus, it is seen that the potential performance of a gyroscopically stabilized platform is very great indeed.

Published performance data[7] indicate that under laboratory conditions the standard-deviation figure $\sigma_{\theta pT}/\sqrt{T_r}$ is of the order of 5.0×10^{-6} rad sec$^{-1/2}$. This means that the realized performance of gyroscopes, at least to the extent that published data are available, is several orders of magnitude below that potentially available from the viewpoint of thermal noise. As electronics engineers, we know that radio receivers and galvanometers[8] are thermally limited, and it seems natural to seek similar performance from other electrical and electromechanical instruments such as gyroscopes.

INFLUENCE OF OTHER FLUCTUATIONS

The large discrepancy between the published performance and the potentially-available performance, if thermal fluctuations are taken to be the ultimate limitation, suggests that other influences may be sources of the much larger fluctuations which are apparently limiting existing gyroscopes. In discussion of other possible sources, vehicle maneuver angles θ_c and platform torque disturbances M_d are first eliminated from consideration, since the vehicle angular velocity $\dot{\theta}_c(t)$ and the torque disturbances must be characterized by power-density spectra which are finite at zero frequency. This follows from the fact that the stochastic components of these signals, which are the only components being considered, have no dc terms and also do not wander off from zero like integrated noise signals. From the transmission functions, (1), it is noted that $W_d(0)$ is zero and that $s^{-1}W_c(s)$, for $s=0$, is also zero. This means that, from Fig. 6, the power-density spectra of the components of the platform velocity caused by these signals is zero at zero frequency and, therefore, there can be no long-term growth of the standard deviation of the platform position from these sources.

Returning to gimbal-torque fluctuations, it would appear that there are a number of possible sources in addition to thermal noise. One is the control torque generator and associated equipment. Although this may be important in present gyroscopes, there is no theoretical reason why fluctuations from this source cannot be made comparable to the thermal-noise fluctuations in the damping fluid. Therefore, this source will not be discussed.

Another source is the interaction of acceleration forces (including gravity) with center-of-mass wander of the rotor caused by bearing noise. Almost all of the gyroscopes manufactured today employ ball-bearings for supporting the rotor. The aspect of bearing noise that affects gyroscope performance is not the acoustical one but rather the lower-frequency wander of spin-axis location. In the presence of gravity and acceleration fields, stochastic movement of the rotor center along the spin axis produces random moments that result in a random drift of the platform.

In order to arrive at an order of magnitude for the effect of bearing noise, information on the low-frequency wander is needed. A limited amount of test data available to the writer indicate that the power-density spectrum of the axial displacement of precision ball bearings in the sizes used in gyroscopes is relatively flat near zero frequency. Furthermore, the amplitude η_b^2 of the spectrum in this region is of the order of

$$\left[\frac{10 \text{ m}\mu}{(\text{rad/sec})^{1/2}}\right]^2.$$

The gimbal moment caused by axial displacement of the rotor is

$$M_{fb} = m_r a_z \alpha_b \tag{10}$$

where m_r is the rotor mass, a_z is the acceleration-gravity component that is normal to the spin and gimbal axes, and α_b is the axial displacement of rotor caused by the bearings. From Fig. 6 and the data given above, the zero-frequency value of the spectrum $\Phi_{\Omega pb}$ of the platform rate caused by bearing noise is

$$\Phi_{\Omega pb}(0) = (m_r a_z \eta_b)^2 W_g{}^2(0). \tag{11}$$

Using (1) and (5) together with (10), we get

$$\frac{\sigma_{\theta pb}}{\sqrt{T_r}} = (2\pi)^{1/2}\left(\frac{m_r a_z \eta_b}{H}\right). \tag{12}$$

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[7] The Kearfott Company, Inc., "Floated rate integrating gyros," *Control Engrg.*, vol. 6, p. 134; October, 1959. This advertisement gives standard-deviation data for Kearfott Model M2516-01A-A gyroscopes. Knowing that the standard-deviation data are based on readings approximately eight minutes apart permits one to arrive at the estimated figure cited above. The data of Table I approximately correspond to the parameters of this gyroscope.

The Sperry Gyroscope Company's SYG-500 floated integrating gyroscope is described in its Publication Number CA-60-0001 dated November, 1959. The published performance curves for this instrument, which has five times the angular momentum of the Kearfott instrument, correspond to a standard deviation figure that is substantially comparable to that quoted in text.

[8] G. Ising, "A natural limit for the sensibility of galvanometers," *Phil. Mag.*, vol. 1, pp. 827–834; April, 1926.

For a 1-*g* acceleration, a gyroscope with the parameters of Table I (next page) is characterized by an acceleration-bearing-noise standard deviation of

$$\frac{\sigma_{\theta pb}}{\sqrt{T_r}} = \frac{(2\pi)^{1/2}25(980)1 \times 10^{-6}}{1.0 \times 10^5} = 0.614 \times 10^{-6} \frac{\text{rad}}{\text{sec}^{1/2}} \cdot \qquad (13)$$

This figure is a factor of ten less than the standard-deviation figure inferred from published performance data for a gyroscope with similar parameters. Thus, bearing noise alone does not fully explain the published performance but it comes much closer to doing so than does thermal noise. It should be noted that in space flight accelerations will be near zero, causing bearing-noise fluctuations associated with acceleration to be much reduced.

In addition to the axial-wander phenomena, bearing noise may produce randomness of the platform angle through other mechanisms, either singly or in combination. Dynamic unbalance, fluctuating gyroscopic torques, and bearing nonlinearities are examples. It is recommended that future studies of gyroscopically-stabilized platforms include analyses of these effects with the objective of improving the correlation between theory and observed behavior.

DISCUSSION

A gyroscopically-stabilized platform has been shown to drift away from its desired angular orientation even though all systematic errors have been eliminated or compensated for. This drift is random in nature and is caused by randomness of platform angular rate. Angular-rate randomness presumably is the result of many influences, but the most significant ones are those that produce power-density spectra of the platform rate that are non-zero for zero frequency. Such influences produce random errors in platform position that increase with the square root of time. Analysis shows that, for a platform stabilized by an integrating gyroscope, platform-torque disturbances and vehicle angular motions are not significant in the build-up of platform error. However, certain random components of gyroscope gimbal torque do cause error build-up. In particular, the error build-up has been quantitatively related to torques produced by thermal fluctuations (Brownian movement) in the damping fluid and to torques produced by acceleration and gravity acting on center-of-mass shifts caused by spin-axis bearing noise.

The error build-up caused by thermal fluctuations in the damping fluid is small. On the basis of parameter values for a medium-sized gyroscope, more than 100 centuries would go by before the platform-error standard deviation would exceed one minute of arc if thermal fluctuations were the only source of error. Published data for gyroscopes in current production imply error performance that is inferior to an ideal thermally-limited gyroscope by about four orders of magnitude (or by eight orders of magnitude if the time to build up a prescribed standard deviation is the criterion).

The error build-up caused by mass shifts induced by ball-bearing noise is quite large when the earth's gravitational field or an equivalent acceleration is acting on the instrument. However, this source of error does not fully explain the published test data referred to above. It is therefore concluded that additional explanations for these data should be sought.

The major conclusion of this paper is that no fundamental barrier to a vast improvement in stable platform performance exists if the same phenomenon that limits the sensitivity of radio receivers and galvanometers also proves to be the basic limitation on gyroscope performance. This phenomenon is thermal noise and, so far, inertial-guidance engineers have not been limited by it because other, less fundamental, sources of fluctuations have been so much larger. Stable platforms that are used on space vehicles may perform somewhat better than those used on earth-bound vehicles, because the effect of one of these other sources will be greatly reduced in view of the diminished influence of acceleration and gravity.

It should not be concluded from this paper that it is possible to achieve the levels of performance that are indicated by the thermal-noise limitation either by existing approaches or by any new scheme that may be devised in the future. The requirements on systematic error elimination become increasingly severe as the random errors are reduced. To make a thermally-limited platform of practical use requires the systematic component of the platform-rate error to be of the order of one second of arc per century, for the example given above.

On the other hand, it is reasonable to conclude from this paper that a continued effort to improve stable-platform performance is justified. Such an effort should be twofold. First, the improvement of existing gyroscopes and associated equipment should be continued. Incorporation of hydrodynamic (gas) bearings is an example of one step in this direction. Second, the search for new ideas for obtaining a stable angular reference should be pressed forward with vigor. Each of these new ideas should be analyzed theoretically, and those which look promising should be put to epxerimental test.

APPENDIX I

NOMENCLATURE

a_z = effective acceleration (including gravity) normal to gimbal and spin axes
e = base of natural logarithms
f_g = viscous-damping coefficient of gimbal
f_p = platform-servomotor damping
H = angular momentum of gyro rotor
J_g = gimbal moment of inertia
J_p = platform moment of inertia
J_r = moment of inertia of rotor

j = square root of minus one
K = Boltzmann's constant
K_v = velocity constant of platform servomechanisms
k_p = platform-servomechanism gain (torque per unit angle of gimbal tilt)
M_c = control torque applied to gimbal axis
M_d = platform-torque disturbance
M_f = torque fluctuation about gimbal axis
m_r = mass of gyro rotor
s = complex frequency used in Fourier transforms
T = absolute temperature
T_r = time to platform position reset
t = time
$W(s, T_r)$ = filter time-varying transmission function
$w(t)$ = filter-weighting function
x = input of filter
y = output of filter
α_b = axial displacement of rotor caused by bearing noise
$\eta_b{}^2$ = amplitude, at zero frequency, of the power-density spectrum of rotor axial displacement caused by bearing noise
θ_v = vehicle angle with respect to inertial space
θ_g = gimbal-tilt angle
θ_p = platform angle with respect to inertial space
σ = standard deviation of filter output
σ_{θ_p} = standard deviation of platform angle θ_p
$\Phi_{xx}(s)$ = power-density spectrum of input to filter
$\Phi_{\Omega_p}(s)$ = power-density spectrum of platform angular rate
Ω_p = platform angular rate with respect to inertial space
ω_r = wheel-spin angular velocity.

APPENDIX II

DERIVATION OF (5) FOR $\sigma(T_r)$

The weighting function $w(t)$ of Fig. 7 that corresponds to an integrator is

$$\left.\begin{aligned} w(t) &= 1 \qquad \text{for } t \geq 0 \\ w(t) &= 0 \qquad \text{for } t < 0 \end{aligned}\right\} . \tag{14}$$

Subsituting this into (4) yields

$$W(s, T_r) = \frac{1 - e^{-sT_r}}{s} . \tag{15}$$

Placing this time-varying transmission function into (3) gives

$$\sigma^2(T_r) = \frac{1}{2\pi j} \int_{-j\infty}^{j\infty} ds F(s, T_r) \tag{16}$$

where

$$F(s, T_r) = \lim_{\epsilon \to 0} 2\pi \left[\frac{e^{+sT_r} - 2 - e^{-sT_r}}{(s - \epsilon)(s + \epsilon)} \right] \Phi_{xx}(s). \tag{17}$$

The splitting of the double-order pole at $s=0$ into two first-order poles at $\pm\epsilon$ is justified by regarding the integration that gives rise to this pole as the limit of the weighting function

$$\left.\begin{aligned} w(t) &= e^{-\epsilon t} \qquad \text{for } t \geq 0 \\ w(t) &= 0 \qquad \text{for } t < 0 \end{aligned}\right\} \tag{18}$$

that is reached when $\epsilon = 0$.

A line integral of the type (16) can be evaluated by separating the integrand into two components:

$$F_1(s, T_r) = \lim_{\epsilon \to 0} 2\pi \left[\frac{e^{sT_r} - 1}{(s - \epsilon)(s + \epsilon)} \right] \Phi_{xx}(s)$$

and

$$F_2(s, T_r) = \lim_{\epsilon \to 0} 2\pi \left[\frac{e^{-sT_r} - 1}{(s - \epsilon)(s + \epsilon)} \right] \Phi_{xx}(s). \tag{19}$$

The contribution of each component is evaluated separately by contour integrals consisting of the imaginary axis and large semicircles centered on the origin. Since T_r is positive and real, a semicircle in the left half-plane is used for F_1 and in the right half-plane for F_2. This choice of contours ensures, in each instance, that the contribution to the integral along the semicircle approaches zero as the radius approaches infinity. Then, use of the residue theorem shows that

$$\sigma^2(T_r) = \sum_m K_{1m} - \sum_n K_{2n} \tag{20}$$

where K_{1m} is the residue of F_1 at the pole at s_m in the left half-plane and K_{2n} is the residue of F_2 at the pole at s_n in the right half-plane. Each summation includes all of the poles of a particular function in its indicated half-plane.

The residues associated with poles of $\Phi_{xx}(s)$ contribute terms of the type $c(1 - e^{-\alpha T_r})$ where c is a constant. These are bounded as T_r approaches infinity. The only unbounded terms arise from the poles at $\pm\epsilon$. Let the residues at these poles be indicated by $K_{1\epsilon}$ and $K_{2\epsilon}$ respectively for the functions F_1 (pole at $-\epsilon$) and F_2 (pole at $+\epsilon$). Then,

$$\lim_{T_r \to \infty} [\sigma^2(T_r)] = \lim_{\epsilon \to 0} [K_{1\epsilon} - K_{2\epsilon}]. \tag{21}$$

But

$$K_{1\epsilon} = 2\pi \left[\frac{1 - e^{-\epsilon T_r}}{2\epsilon} \right] \Phi_{xx}(-\epsilon)$$

$$-K_{2\epsilon} = 2\pi \left[\frac{1 - e^{-\epsilon T_r}}{2\epsilon} \right] \Phi_{xx}(\epsilon). \tag{22}$$

Substituting these values into (20) and taking the limit as ϵ goes to zero yields (5).[9]

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[9] W. B. Davenport, Jr. and W. L. Root, "An Introduction to the Theory of Random Signals and Noise," McGraw-Hill Book Co., Inc., New York, N. Y., p. 69, (4–83); 1958. This source gives an equivalent expression to (5) in the time domain.

The Astronautic Chart*

ROY C. SPENCER†, FELLOW, IRE

***Summary*—The Astronautic Chart is a nomograph or alignment chart so arranged that a single straight line marks off values of the velocity, mass, mean distance, period, and acceleration of any two-body orbiting system. It is illustrated with numerous examples of orbits of planets about the sun, moons about their planets, and artificial earth satellites.**

All scales give correct values at the extremities of the minor diameter of the elliptical orbit. In the case of binary stars where the masses are comparable, the scales also give correct values of the *total mass*, *total separation*, *relative velocity*, and *relative acceleration*.

INTRODUCTION

LAUNCHING of the twenty-two artificial earth satellites and lunar and space probes during the first two years of the space age has again focused attention of students and scientists on the laws of Kepler and Newton that control the paths of the planets around the sun and of natural satellites around their planets. These laws apply as well to the motions of artificial satellites.

The specialist in celestial mechanics or space travel deals with orbit problems that become very intricate when precise answers are needed. But a general understanding of the physical laws and their application is much easier to attain. In fact, with the aid of the Astronautic Chart[1–4] inserted between the pages of this issue, the would-be astronaut can quickly obtain an appreciation of the motions within the solar system. This simple graphical device shows at a glance the interrelations among distances, orbital periods, and masses, as well as velocities and accelerations.

The diagram is a nomograph or alignment chart, so arranged that a single oblique line gives the characteristics of a particular orbit. The oblique lines illustrate numerous systems of bodies in orbital motion; other lines may be drawn to solve problems concerning possible future satellites, or asteroid orbits.

Thus, in Fig. 1 the five vertical scales indicate the following, from left to right.

V is the mean orbital velocity, in kilometers per second. For a circular orbit, this is the constant speed of the moving body. If the orbit is an ellipse, V is the geometric mean (square root of the product) of the velocities at perihelion and aphelion, or at perigee and apogee.

M is the mass of the central body, in terms of the sun's mass as a unit. Strictly speaking, this is the combined mass of the two bodies, but for graphical purposes the mass of any planet or satellite can be regarded as negligible compared to the mass of its primary. (Only our moon has a mass as great as 1/80 of its central body.)

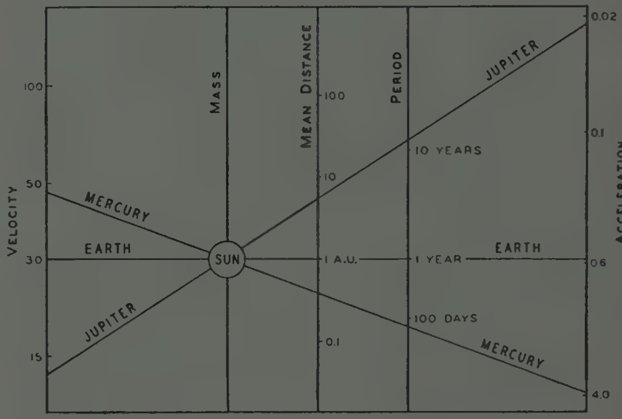

Fig. 1—Simplified portion of the Astronautic Chart. The earth-sun solution is horizontal.

a is the mean distance from the primary body, expressed in astronomical units or kilometers. In the case of an elliptical planetary orbit, this is the arithmetic mean (or average) of the perihelion and aphelion distances. One astronomical unit (A.U.) is the mean distance of the earth from the sun, about 93 million miles.

P is the period of orbital revolution, expressed in years at the top. This scale is successively changed to days, hours, and minutes to fit the short periods in the lower part of the chart. Since the mean angular velocity, $d\theta/dt$, is 360° divided by the period, we could construct such a scale collinear with the period scale. For example, a period of 90 minutes yields a mean angular velocity of 4° per minute.

g is the "mean" acceleration toward the central body, expressed in centimeters per second each second. For example, in the case of Jupiter, this is the velocity of free fall toward the sun, one second after starting from rest, of a body at Jupiter's distance from the sun.

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* Original manuscript received by the IRE, December 23, 1959. Revised from a paper published in *Sky and Telescope*, vol. XVII, no. 11; September, 1958.

† The Martin Co., Baltimore Div., Baltimore, Md. Formerly with Sylvania Electric Products Inc., Waltham, Mass.

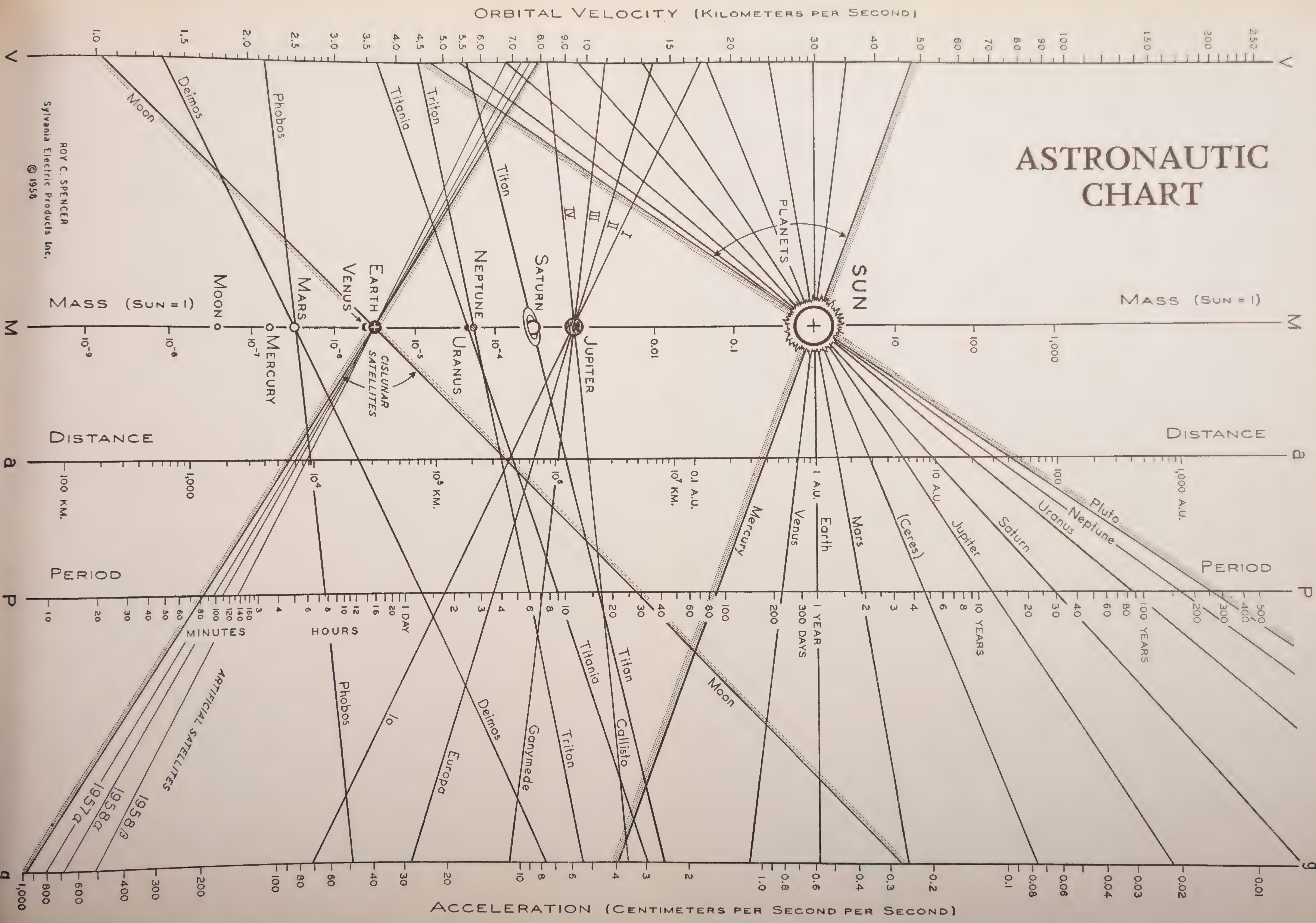
[1] R. C. Spencer, "The Astronautic Chart for Planets and Satellites," Sylvania Electric Products Inc., Waltham Labs., Waltham, Mass., Rept. No. A58-7B; March, 1958.

[2] R. C. Spencer, "Derivation of Formulas for the Astronautic Chart," Sylvania Electric Products Inc., Waltham Labs., Waltham, Mass., Rept. No. A58-7C; June, 1958.

[3] R. C. Spencer, "Fundamental Constants and Tables for Earth Satellites," Sylvania Electric Products Inc., Waltham Labs., Waltham, Mass., Rept. No. A58-7C; September, 1958.

[4] Numerical values of astrophysical constants used in references 1–3 were obtained from C. W. Allen, "Astrophysical Quantities," University of London, Athlone Press, London, Eng.; 1955.

ASTRONAUTIC CHART



ROY C. SPENCER
Sylvania Electric Products Inc.
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USING THE CHART

As an example of the use of the Astronautic Chart, consider the planet Saturn, which moves around the sun in a period of about $29\frac{1}{2}$ years. Lay a straightedge across the chart, passing through the points marked "sun" on the $\boldsymbol{M}$ scale and just below 30 years on the $\boldsymbol{P}$ scale. Extended to the left, this straight line shows that Saturn's orbital velocity is some $9\frac{1}{2}$ kilometers per second. From the intersection with the middle scale, the mean distance of Saturn from the sun is about 9.5 astronomical units, and the right-hand scale shows that the planet's acceleration toward the sun is much less than 0.01 cm/second^2.

There are corresponding lines drawn through the sun for the other principal planets, from Mercury to Pluto, as well as for a representative asteroid, Ceres.

In the lower part of the chart, each planet is indicated at the proper place on the mass scale. Through the planet symbols are drawn straight lines giving chart solutions for some of their satellites. The line for Jupiter's brightest moon, III (Ganymede), provides the following information: orbital velocity, about 11 kilometers per second; distance from Jupiter, over a million kilometers; period, seven days; and acceleration toward the planet 11 cm/second^2.

There is quite an overlap for some members of the solar system in both acceleration and velocity. For instance, the acceleration of our moon toward the earth approximates that of Mars toward the sun. Also, the velocities of the earth's present artificial satellites are about the same as the orbital velocity of Uranus around the sun.

ARTIFICIAL SATELLITES

An important use of the chart is the solution of problems involving earth satellites, whether natural or artificial. Two shaded lines run diagonally through the earth symbol on the mass axis, one for the moon, the other for a hypothetical satellite that would just graze the earth's surface. The double triangle bounded by these lines is the chart area including all possible earth satellites closer to us than the moon, that is, the *cislunar* area.

Three actual satellites have been plotted, 1957α, 1958α and 1958β, showing the range of initial period of the earlier artificial satellites. A typical moon shot, consisting of a narrow elliptical orbit reaching out to the moon, would be indicated by a line from the earth to a point marking approximately half the moon's distance. The solution for the proposed "one-day" satellite, which will hover over a particular point on the equator, will be a line through the one-day period.

Space-travel enthusiasts who expect that some day man-made satellites will be circling other planets to observe them close at hand will find use for the chart of Fig. 2, where minimum satellites have been plotted for all planets except Pluto and Venus. Each such satellite is presumed to circle its primary just above the surface we see, whether that surface is solid, as on Mercury and Mars, or gaseous, as on Jupiter and Saturn.

The minimum-satellite velocities differ widely, yet their periods are of the same order, ranging from over four hours for a minimum satellite of Saturn to 84.5 minutes for one of the earth. This is because the period for a spherical planet depends only on the planet's density. All spheres with the earth's density would have minimum-satellite periods of 84.5 minutes. A sphere of ice, whether 10 miles in diameter or the size of Jupiter, would have a period of about three hours and 27 minutes for such a grazing moonlet. This raises the question, "Is the interior of Jupiter made of ice?"

On the full-sized Astronautic Chart, the extension of any minimum-satellite line to intersect the acceleration ($\boldsymbol{g}$) scale indicates the surface gravity of the planet: for example, 980 cm/second^2 for the earth. (This approximation would be exact if the planet were spherical and nonrotating.)

Attempts will eventually be made to establish a *circumlunar satellite* orbit by slowing down the velocity of a rocket which is about to miss the moon's surface. The Astronautic Chart also serves to answer numerical questions about the orbits of such artificial satellites concerning the natural satellites of planets. As an example, a line has been drawn in the minimum-satellite chart for a minimum circumlunar vehicle. From this we see that a rocket ship just clearing the lunar surface would require about 110 minutes to make a trip around the moon. The reader may easily find the orbital velocity and period of a rocket revolving around the moon at a distance above its surface of, say, 900 kilometers (2638 kilometers from its center).

KEPLER'S AND NEWTON'S LAWS

The three middle scales for mass, mean distance, and period provide a solution for Kepler's third or harmonic law, which holds for elliptical as well as circular orbits:

$$\frac{a^3}{P^2} = \frac{GM}{4\pi^2} \tag{1a}$$

in which G is the constant of gravitation.

If the mass M were uniformly distributed throughout a sphere of radius a and density ρ, then $M=(4/3)\pi\rho a^3$. On substituting this in (1a), we obtain

$$\rho P^2 = 3\pi/G, \tag{1b}$$

indicating, as was mentioned earlier, that the mean density depends only on the period.

Design of the chart is facilitated by expressing the variables in (1a) in terms of the $\boldsymbol{V}$ and $\boldsymbol{g}$ that appear at

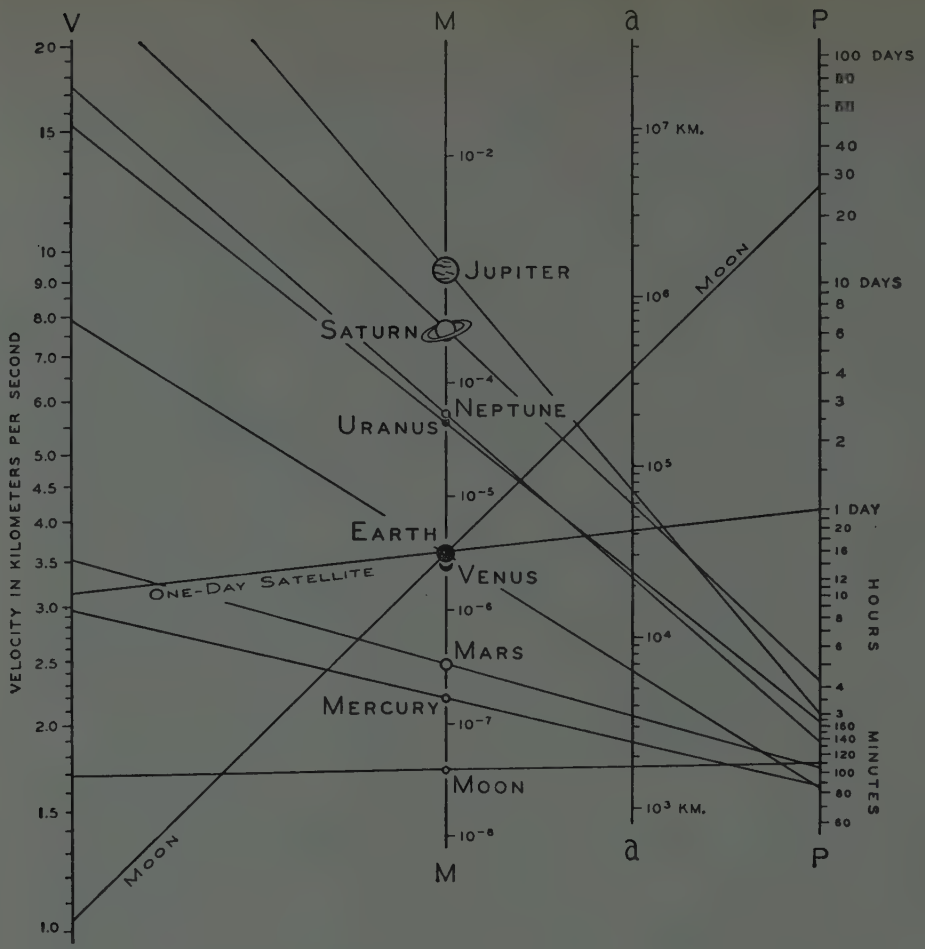

Fig. 2—Minimum satellite orbits of planets (and moon). Orbits of the one-day earth satellite and the moon have been included.

the sides of the chart. One writes the expressions for velocity and centripetal acceleration in the case of uniform circular motion discussed in elementary physics:

$$V = 2\pi a/P; \qquad g = V^2/a.$$

By rearrangement,

$$a = V^2/g; \qquad P/2\pi = a/V = V/g. \qquad (2,3)$$

Newton's Law of Universal Gravitation states that $g = GM/a^2$, whence

$$GM = a^2 g = V^4/g. \qquad (4)$$

Note that (2)-(4) are of the form V^n/g.

Construction of the Chart

Many readers are familiar with the multiplication nomograph or alignment chart for an equation of the form $z = xy$. Thus, in the simplest form, x and y are parallel logarithmic scales with the same logarithmic base, while the scale for z is placed *midway* between them with a logarithmic base *half as great*.

Actually, there is some flexibility in the choice of the logarithmic bases, as shown in the diagram of Fig. 3 for the case of $z = x^n y$. Choose any convenient base A for the cycle from 1 to 10 of the x scale. Then divide by n as shown on the left. On the right, any convenient parallel base B is shown for scale y. The intersection of the diagonals of the parallel line segments A/n and B are then marked off to find the *horizontal position* and *length* of the base C for z.

It can be shown from the construction that the position of the z scale is determined by

$$D = A/(A + nB) \qquad (5)$$

where D is the fraction of the distance from the left-hand scale to the right-hand one. Furthermore,

$$C = AB/(A + nB), \qquad (6)$$

where C is the base length (height) of the z scale.

These expressions can now be applied to (2), (3), and (4), each of which is of the form $z = V^n \times 1/g$. For convenience, let the logarithmic base for V, the left-hand scale of the Astronautic Chart, be of unit length, repre-

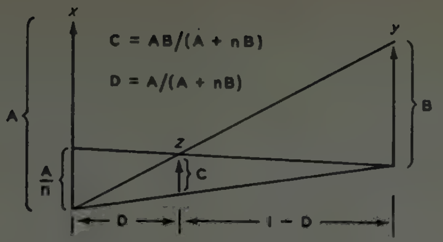

Fig. 3—Nomograph for $z=x^ny$, with C, A, and B the respective logarithmic scales.

sented by 1. For the right-hand scale, we adopt a base half as long, represented by $\frac{1}{2}$. Our choices make $A=1$ and $B=\frac{1}{2}$.

In (4), (2), and (3), n has the values 4, 2, and 1, for the ***M***, ***a***, and ***P*** scales, respectively. Substituting these values in the expressions above for D and C gives Table I for use in Fig. 4.

TABLE I

Scale	*V*	*M*	*a*	*P*	1/*g*
Position	0	1/3	1/2	2/3	1
Base	1	1/6	1/4	1/3	1/2

The reader who wishes to subdivide further the scales of the Astronautic Chart will find the proper values in any scale of logarithms to the base 10.

Further information about nomograms may be found in such texts as Douglass and Adams,[5] and Mackey.[6]

Extension to Elliptic and Binary Orbits

The three middle scales naturally apply to elliptic orbits because they solve Kepler's third law. However, after including in the chart the additional scales for ***V*** and ***g*** on the basis of circular orbits, the author was agreeably surprised to find that they were also exact at the extremities of the minor diameter of the ellipse. Thus, in Fig. 5, the circular orbit is "equivalent" to the elliptical orbit in period and major diameter. The orbits intersect at the extremities of the minor diameter BB′ of the ellipse, at which points the accelerations are equal and the velocities are the same in magnitude, whence the term "circular velocity."

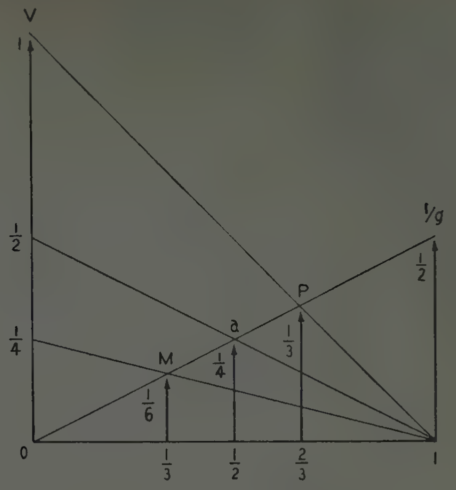

Fig. 4—Construction of the Astronautic Chart.

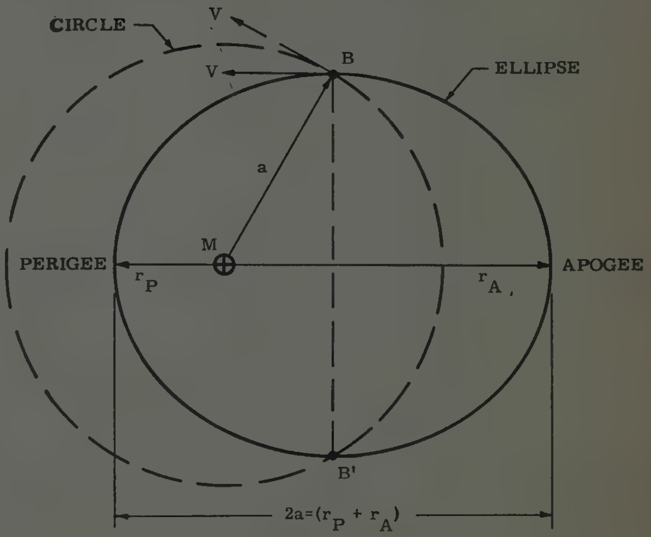

Fig. 5—Equivalent circular and elliptic orbits.

Newton pointed out that, in a binary system consisting of two masses m_1 and m_2 revolving about their common center of gravity, Kepler's law still holds if *M* is the *sum* of the masses and *a* is the *mean separation*. It now appears that the ***V*** and ***g*** scales also hold at the extremities of the minor diameter of the *relative* ellipse, providing they represent *sums* of corresponding scalar magnitudes (or vector differences).

Originally the author had inserted several orbits of binary stars, such as Capella and Sirius, but later he decided to omit them in order to simplify the chart and restrict its illustrations to the solar system. But whether the reader is interested in earth satellites, interplanetary exploration, double stars, or a combination of these, the Astronautic Chart can be used to study them further.

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[5] R. D. Douglass and D. P. Adams, "Elements of Nomography," McGraw-Hill Book Co., Inc., New York, N. Y.; 1947.

[6] C. O. Mackey, "Graphical Solutions," John Wiley and Sons, Inc., New York, N. Y.; 1936.

A Study of Natural Electromagnetic Phenomena for Space Navigation*

R. G. FRANKLIN† AND D. L. BIRX†, MEMBER, IRE

Summary—**A study has been carried out for the United States Air Force investigating the use of natural electromagnetic radiation in the space environment for navigational purposes. Radiations from the sun, the stars, and interstellar space in both the visible and RF portions of the spectrum and also cosmic rays have been investigated.**

Emphasis has been placed on the measurement of velocity in space utilizing the Doppler phenomenon. Equipment and techniques useful in deriving velocity information from Doppler shift measurements are described and figures for expected accuracy are derived. Other passive techniques having possible application to space navigation such as the measurement of total solar radiation and solar diameter are briefly discussed.

INTRODUCTION

THE Franklin Institute has been engaged in a study of natural electromagnetic radiation for a period of approximately one year. Radiations from the sun, the stars and interstellar space in both the optical and radio frequency portions of the spectrum have been investigated, with the aim of determining their value to space navigation, particularly in the measurement of velocity utilizing the Doppler phenomenon. In this paper, sources and techniques useful in deriving velocity information from Doppler shift measurements are described and figures for expected accuracy are derived. These data are then compared in accuracy with results of two other passive techniques, the measurement of total solar radiation and of solar diameter.

Whenever there is relative motion between a source of radiation and an observer, the radiation observed is shifted in wavelength an amount $\Delta\lambda$, which is proportional to the relative velocity according to the equation

$$\Delta\lambda = \lambda_1 - \lambda_2 = \frac{v\lambda_1}{c}, \tag{1}$$

where

v = relative velocity of source with respect to observer,
c = velocity of light,
λ_1 = wavelength of radiation observed with no relative motion,
λ_2 = wavelength observed with relative motion.

This equation permits the measurement of the radial velocity of a space vehicle with respect to a source of radiation, provided that λ_1 is known, and $\Delta\lambda$ can be measured. The velocity error δv depends on the accuracy of the measurement of λ_1 and λ_2 and is given by the formula

$$\delta v = \frac{2c\delta\lambda}{\lambda_1}, \tag{2}$$

where $\delta\lambda$ = error in the determination of wavelength. The suitability of Doppler methods of velocity determination therefore depends on the magnitude of $\delta\lambda$ which, in turn, is influenced by the intensity and shape of the spectral lines and the measuring equipment characteristics. As for the lines themselves, the most desirable characteristics are high intensity and good sharpness (narrow width).

CHARACTERISTICS OF SPECTRAL LINES IN THE OPTICAL REGION

Aller[1] has outlined the causes of broadening of spectral lines. The most important of these for our present discussion is the Doppler effect, whereby lines are broadened due to the random kinetic motion of the atoms, the turbulence of large masses of gases in the atmospheres of the heavenly sources, and the axial spin of the source if the light is integrated from all parts of the source. Aller derived an expression for the halfwidth of the line, broadened only by Doppler effects:

$$\Delta\lambda = 7.16 \times 10^{-7}\lambda\sqrt{\frac{T}{\mu}}, \tag{3}$$

where

$\Delta\lambda$ = the halfwidth of the line,
λ = the wavelength of the line,
T = the temperature (°K),
μ = the molecular weight of the radiator (or absorber).

The lines are broader, therefore, for longer wavelengths and higher temperatures, and for the lighter elements.

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† Labs. for Research and Development, The Franklin Institute, Philadelphia, Pa.

[1] L. H. Aller, "Astrophysics," The Ronald Press Co., New York, N. Y., ch. 8; 1953.

The shape of the spectral lines at the center is largely determined by Doppler effects, whereas other causes, *e.g.*, radiation and collision damping, fix the shape at greater distances from the center of the line. Analysis of the Doppler effects shows that even narrow lines have comparatively flat tops. Therefore, for wavelength determination, the sides of the line should be utilized wherever possible.

With regard to intensity, the sun is the only source which provides adequate energy for high-accuracy spectroscopic measurements. The Babcock magnetograph,[2] for example, is able to provide an equivalent velocity resolution of one meter per second in Doppler measurements, using the sun as the source. The radiation received from even the brightest stars, however, is approximately only 1×10^{-10} as intense. Stars differ greatly on the basis of both apparent intensity and spectra. They may be classified according to their spectra which in turn are a function of the star temperature. Table I lists the spectrum classes of the main sequence stars, the associated temperatures and some of the more characteristic features of each class. Each class is further subdivided, as B0, B1, $\cdots$, B9.

By assuming that the sun and the stars radiate as black bodies, the fraction of the total emittance over any given wavelength range can be calculated from the radiation laws

1)
$$W_{bb} = \sigma T^4, \tag{4}$$

where

W_{bb} = the total radiant emittance of a black body at the temperature T,

T = absolute temperature (°K),

σ = the Stefan-Boltzmann constant;

2)
$$\frac{dW_{bb}}{\Delta\lambda} = W_\lambda = \frac{c_1\lambda^{-5}}{e^{c_2/\lambda T} - 1}, \tag{5}$$

where

W_λ = spectral emittance at the wavelength λ over a bandwidth $\Delta\lambda$,

c_1, c_2 = first and second radiation constants.

From this, the spectral irradiance $H_{s\lambda}$ over the wavelength band $\Delta\lambda$ of a source is given by

$$H_{s\lambda} = \frac{W_\lambda}{W_{bb}} \times H_s, \tag{6}$$

where H_s is the total irradiance of the source on the earth.

In the case of the sun, H_s, the solar constant is 0.14 watt/cm², and for a star of zero bolometric magnitude, H_s has been determined to be 2.27×10^{-12} watts/cm².

TABLE I

SPECTRUM CLASSES AND THEIR CHARACTERISTICS

Spectrum Class	Mean Color Temperature	Characteristic Features
0	23,000°K	Faint He lines present.
B0	21,000	He lines intense, reaching a maximum in B2 stars and fading out in B9 stars.
A0	11,000	*H* lines reach a maximum; metal lines begin to appear.
F0	7,300	*H* lines fade out; metal lines, especially Ca (*H* and *K*), increase.
G0	5,900	Metal lines appear in great numbers.
K0	5,280	Metal lines are most prominent; Ca (*H* and *K*) reach a maximum intensity; molecular bands appear.
M0	3,590	Bands of TiO are prominent.
The sun is a G2 star.		

The spectral irradiance over a bandwidth of 1 A has been calculated for various wavelengths from the sun and from stars of zero bolometric magnitude and different spectrum classes. These are given in Table II. Included also are the wavelengths of maximum energy calculated by the relationship

$$\lambda_{\max} = (2.8971/T) \times 10^7 \text{ A}. \tag{7}$$

The last column lists the apparent visual magnitudes.

For stars other than those of zero bolometric magnitude, the irradiances may be calculated from the formula

$$m_2 - m_1 = 2.5 \log \frac{L_1}{L_2},$$

where m_2 and m_1 are the visual magnitudes of the unknown and standard star (of zero bolometric magnitude) of the same spectral class, and L_1 and L_2 are the intensities of the unknown and standard stars respectively. Since the unknown star is being compared to a standard star of the same temperature, L_1 and L_2 may represent their respective irradiances at any wavelength. These data for some of the brightest stars appear in Table III.

While there is much general information as to the lines to be found in stellar spectra, such as the excellent group of photographs published by Morgan, Keenan, and Kellman,[3] there is a scarcity of definite photometric data on the line profiles. Hiltner and Williams[4] have published data for eight stars in their "Photometric Atlas of Stellar Spectra." Fig. 1 is a reproduction of a trace for Rigel in the region of the Ca II *H* and *K* lines at 3968 and 3934 A, and of H_γ and H_δ lines at 4340 A and 4102 A. Fig. 2 shows a similar trace for Sirius. The spectra of cooler stars exhibit many lines, as can be

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[2] H. W. Babcock, "The solar magnetograph," *Astrophys. J.*, vol. 118, pp. 387–396; November, 1953.

[3] W. W. Morgan, P. C. Keenan, and E. Kellman, "An Atlas of Stellar Spectra," University of Chicago Press, Chicago, Ill.; 1943.

[4] W. A. Hiltner and R. Williams, "Photometric Atlas of Stellar Spectra," University of Michigan, Ann Arbor, Mich.; 1946.

TABLE II

SPECTRAL IRRADIANCE OF STARS OF ZERO BOLOMETRIC MAGNITUDE
(WATTS/CM²/A)

Spectral Class	Temperature °K	$H_{s\lambda}$ λ4000A 10^{-16}	$H_{s\lambda}$ λ5000A 10^{-16}	$H_{s\lambda}$ λ1μ 10^{-16}	$H_{s\lambda}$ λ5μ 10^{-19}	$H_{s\lambda}$ λ10μ 10^{-19}	λ_{max} A	Apparent Visual Magnitudes
B0	21,000	1.66	0.84	0.078	0.167	0.011	1,317	2.8
B5	16,000	2.65	1.46	0.156	0.369	0.024	2,069	1.4
A0	11,000	3.70	2.75	0.381	1.064	0.0726	2,708	0.6
A5	8,700	4.19	3.18	0.681	2.10	0.143	3,408	0.2
F0	7,300	3.90	3.41	0.855	3.47	0.241	3,915	0.0
F5	6,400	3.41	3.36	1.053	5.00	0.352	4,457	0.0
G0	5,900	2.84	3.12	1.172	6.26	0.449	4,910	0.5
G5	5,570	2.51	2.90	1.276	7.19	0.518	5,267	0.1
K0	5,280	2.37	2.88	1.362	8.46	0.610	5,912	0.17
K5	4,550	1.63	2.36	1.59	11.90	0.943	6,898	0.6
M0	3,590	0.696	1.40	1.83	23.49	1.826	8,048	1.2
M5	2,800	0.06	0.36	1.63	4.85	0.405	10,350	2.4
G2(sun)	5,760	1.5×10^{11}	1.9×10^{11}	7.6×10^{10}	4.0×10^{11}	2.9×10^{10}	5,030	−26.7

TABLE III

SPECTRAL IRRADIANCE OF SOME BRIGHT STARS
(WATTS/CM²/A)

Star	Class	Visual Magnitudes	$H_{s\lambda}$ λ4000A 10^{-16}	Star	$H_{s\lambda}$ λ5000A 10^{-16}	Star	$H_{s\lambda}$ λ1μ 10^{-16}	Star	$H_{s\lambda}$ λ5μ 10^{-19}	Star	$H_{s\lambda}$ λ10μ 10^{-19}
Sirius	A0	−1.6	28.0	Sirius	21.0	Betelgeuse	3.6	Betelgeuse	41.5	Betelgeuse	3.08
Canopus	F0	−0.9	8.9	Canopus	7.8	Sirius	2.9	Arcturus	8.2	Arcturus	0.6
β Centauri	B1	0.9	8.1	β Centauri	5.1	Canopus	1.9	Sirius	8.1	Sirius	0.6
Vega	A0	0.1	6.7	Vega	4.3	Arcturus	1.3	Canopus	7.9	Canopus	0.55
Rigel	B8	0.3	6.3	Rigel	3.9	Vega	0.6	Vega	1.7	Vega	0.12
Achernar	B5	0.6	5.5	Achernar	3.0	Rigel	0.5	Rigel	1.4	Rigel	0.09
Spica	B2	1.2	4.7	Arcturus	2.8	β Centauri	0.4	Pollux	1.3	Pollux	0.09
Arcturus	K0	0.2	2.3	Spica	2.4	Altair	0.35	Altair	1.1	Altair	0.08
Altair	A5	0.9	2.2	Betelgeuse	2.4	Achernar	0.3	β Centauri	0.9	β Centauri	0.05
Deneb	A2	1.3	1.8	Altair	1.7	Spica	0.23	Achernar	0.8	Achernar	0.05
Fomalhaut	A3	1.3	1.6	Deneb	1.2	Deneb	0.22	Deneb	0.7	Deneb	0.05
Betelgeuse	M2	0.9	0.77	Fomalhaut	1.2	Fomalhaut	0.22	Fomalhaut	0.7	Fomalhaut	0.05
Pollux	K0	1.2	0.37	Pollux	0.45	Pollux	0.21	Spica	0.5	Spica	0.03

seen from Fig. 3 (a) and 3 (b), which are copies of a representative portion of the photometric traces of the spectra of Procyon and Arcturus.

ACCURACY OF VELOCITY DETERMINATION IN THE OPTICAL REGION

As can be seen from Table III, the radiation available from the stars for Doppler measurements is small. An instrument of high resolution and dispersion is necessary to permit accurate measurements, and detectors of great sensitivity are necessary to record small line displacements.

The common technique used in measuring Doppler shift is that of employing two receiving channels separated by a fixed frequency interval and situated on either side of the center of the line to be measured. The outputs of the two channels are compared as the pair is moved across the line, and at the instant a null is obtained (when the outputs of both channels are equal) the center line frequency is deduced as being midway between the two channel frequencies. Any shift in line frequency necessitates a corresponding shift of the two channels in order to reestablish a null. With such a system, it has been shown[5] that, to a good approximation, the change in flux falling on the balanced detector system caused by a displacement $d\lambda$ of an absorption line is given by

$$dL = 2\alpha E_\lambda d\lambda, \tag{8}$$

where

dL = the change in flux,
α = the fraction of radiation absorbed at the center of the line,
E_λ = flux emerging from the spectrograph at λ (without absorption),
$d\lambda$ = the displacement of the line.

The total flux falling on the detectors is

$$L = E_\lambda \Delta\lambda(1 - \alpha/2), \tag{9}$$

where

L = the total flux,
$\Delta\lambda$ = the halfwidth of the line.

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[5] "The Franklin Institute Fourth Quarterly Engineering Report on the Study of Electromagnetic Phenomena for Space Navigation," Philadelphia, Pa., Contract AF 33(616)-5898, Project No. 8(623-5219); June 30, 1959.

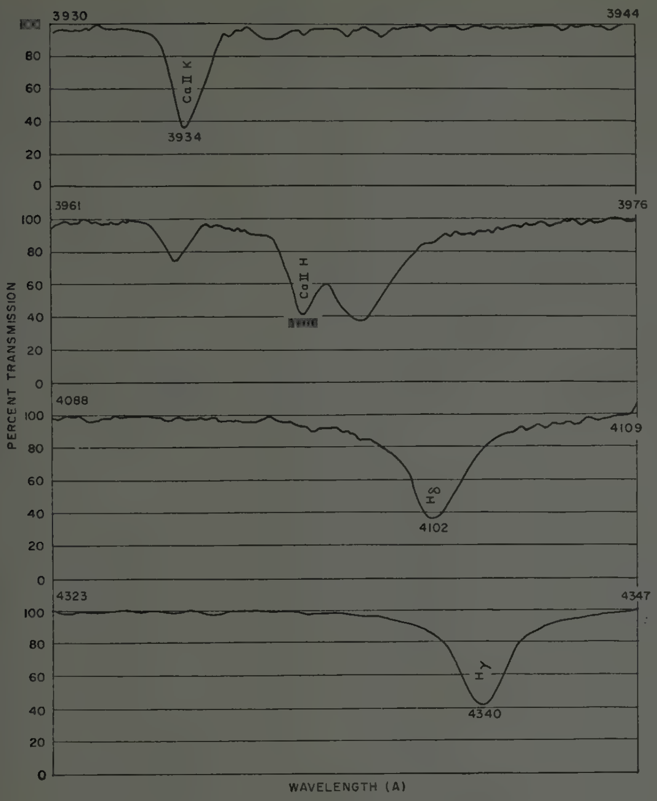


Fig. 1—Photometric traces of spectrum of Rigel.

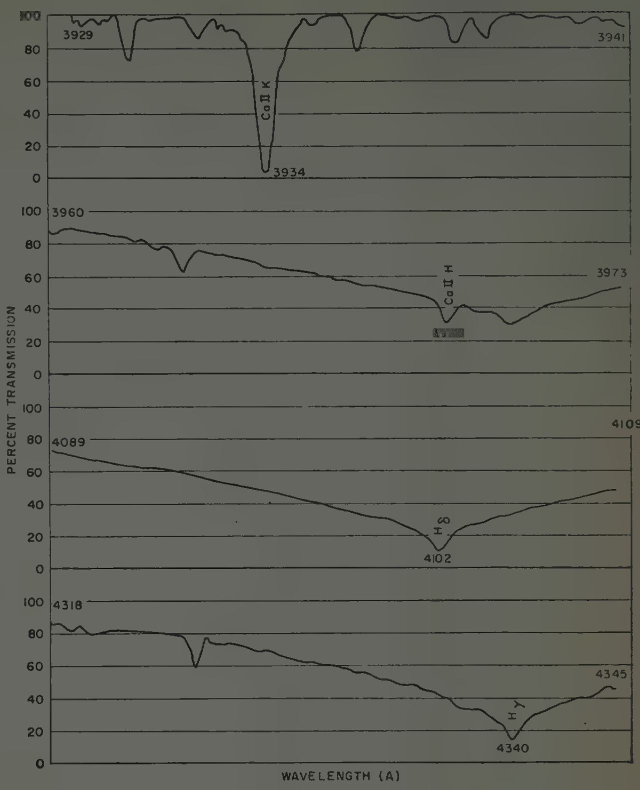
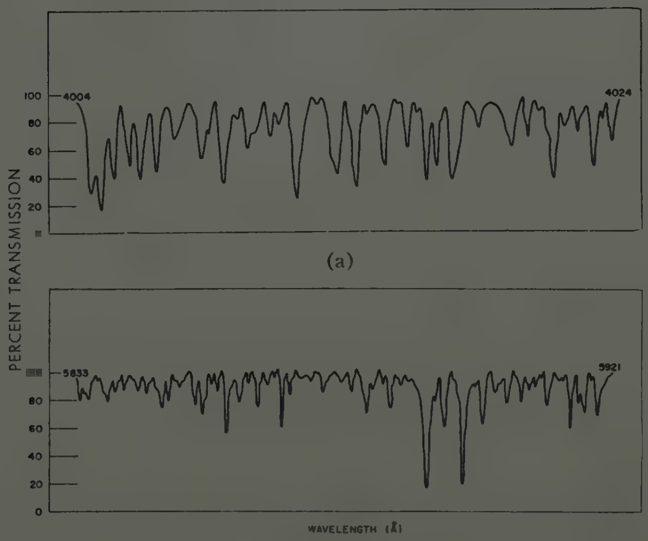


Fig. 2—Photometric traces of spectrum of Sirius.



(a)

(b)

Fig. 3—Photometric traces of stellar spectra.
(a) Procyon. (b) Arcturus.

The ratio of the change in flux to the total flux is therefore

$$\frac{dL}{L} = \frac{2\alpha E_\lambda d\lambda}{E_\lambda \Delta\lambda(1 - \alpha/2)} = \frac{2\alpha d\lambda}{\Delta\lambda(1 - \alpha/2)}. \qquad (10)$$

It is evident from this equation that narrow, strongly absorbing lines are best suited to velocity measurements. The above derivation assumes that the halfwidth of the line and the width of the signal channels are approximately equal.

The equation given below relating the various parameters involved in the detection process has been derived by the workers at the General Precision Laboratory.[6]

$$\sigma(V) = \frac{c}{\lambda}\,\frac{1}{E_\lambda \alpha S}\sqrt{\frac{Ge}{2t}\left[I_D + E_\lambda S\left(W - \frac{\alpha\Delta\lambda}{2}\right)\right]}. \qquad (11)$$

In this equation

$\sigma(V)$ = error in velocity measurements,
c = velocity of light,
λ = wavelength of the light used,
S = spectral sensitivity of the detector tubes,
G = gain of the detector tubes,
e = the charge of an electron,
t = the smoothing time,
I_D = the dark current of the detector tubes,
W = the bandwidth of the signal channel (exit slits of the spectrograph),
$\Delta\lambda$ = halfwidth of the absorption line.

The flux E_λ emerging from the spectrograph system is a function of the irradiance of the source, the area of the collecting telescope, and the attenuation of the optical system. It may be represented by

$$E_\lambda = H_s d\lambda 0 \epsilon, \qquad (12)$$

where

$H_s d\lambda$ = the irradiance of the source,
0 = the area of the telescope collector,
ϵ = the attenuation of the optics.

The errors in the measurement of the velocity $\sigma(V)$ vs the diameter of the collecting telescope were computed by the use of this equation for Sirius and for Rigel, Figs. 4–6, using the parameters indicated in the figures. It may be seen that detectors of the highest quality are essential by comparing Figs. 5 and 6, where computations were made for Rigel using different tube parameters.

Systems for Doppler Measurements

While it is not our present aim to propose a specific design for an instrument, certain features can be outlined and suggestions made on the basis of an examina-

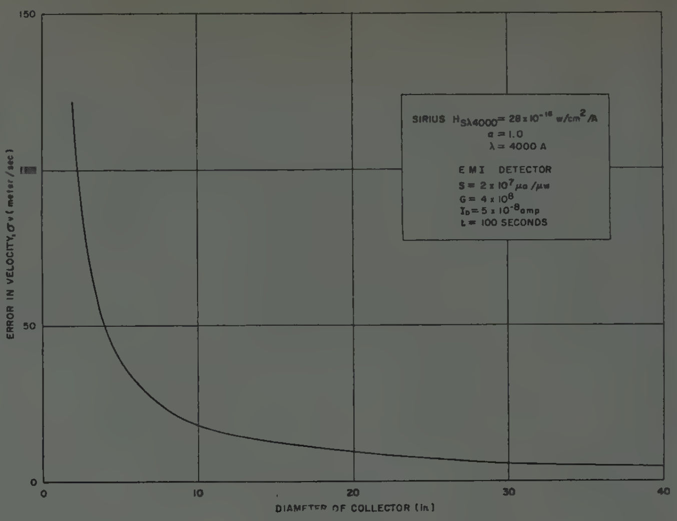


Fig. 4—Variation of errors in velocity measurements with diameter of collector

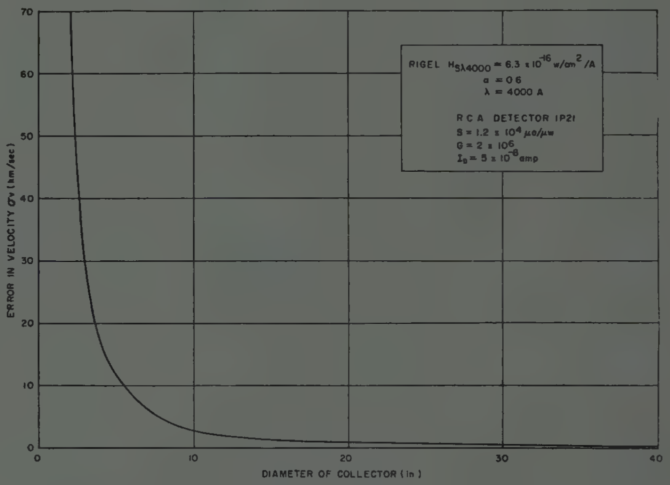


Fig. 5—Variation of errors in velocity measurements with diameter of collector

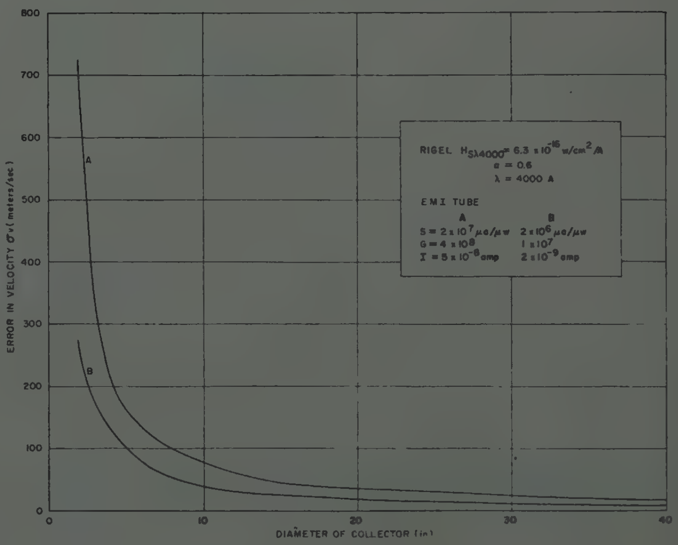


Fig. 6—Variation of errors in velocity measurements with diameter of collector.

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[6] "Interim Engineering Report on the Study of Electromagnetic Phenomena for Space Navigation," General Precision Laboratory Inc., Pleasantville, N. Y., Contract No. AF 33(616)-5487; January, 1959.

tion of the preceding results. Such an instrument should have the following components:

1) A collecting telescope. From Figs. 4–6, it can be seen that little is to be gained by increasing the diameter of the collector above 8 or 10 inches which would therefore be a practical compromise between efficiency and weight.

2) The spectrograph. In order to attain the necessary dispersion and resolution an echelle-prism system might be used to advantage, since such a system combines high dispersion and resolution with efficiency.

3) A slit system in the focal plane of the spectrograph. This slit system would be so constructed as to divide the radiation from a line, permitting the flux from each half to fall on a detector. The output from the two detectors would be balanced at zero velocity.

4) Detectors. As pointed out above, the detectors must be of the highest quality.

Because of the low signal level, it is interesting to examine the effect of using more than one line by employing more than one slit in the detector assembly. The total change in the flux for n lines in the same spectral range would then be n times the average change for one line or

$$dL = 2n\bar{\alpha}\overline{E}_\lambda\overline{d\lambda},$$

and (12) becomes

$$\sigma(V) = \left(\frac{C}{\bar{\lambda}n\bar{\alpha}\overline{E}_\lambda S}\right) \cdot\sqrt{\frac{Ge}{2t}\left[I_D + n\overline{E}_\lambda S\left(W - \frac{\bar{\alpha}\Delta\lambda}{2}\right)\right]}. \qquad (13)$$

For any given star, E_λ and $d\lambda$ will remain essentially constant over a small wavelength range, but α may vary from 1 to 0. Fig. 7 shows the effect of varying α for $H_{s\lambda}=1\times10^{-16}$ watts/cm² A, for different values of n (heavy curves). The other parameters are as indicated in the figure. It can be seen that as α decreases, the velocity error $\sigma(V)$ increases rapidly with a limiting value of ∞ when $\alpha=0$. Thus it is necessary to use only those lines having large values of α, and weakly absorbing lines must be excluded. To illustrate this, assume that the value of α remains constant ($\alpha=0.9$). If one slit is used, $\sigma(V)=153$ meters per second as shown by × on the appropriate curve. If 2 slits are used, $\sigma(V)=96$ meters per second. Similarly, for $n=5$, 10, 20, and 30 slits, the values of $\sigma(V)$ are 56, 37, 25, and 20 meters per second, as shown by the points marked × in the figure.

But if α is also varied when several slits are used, the velocity errors may be increased. Assume the following values of α for the various values of n:

n	nH_s (*watt/cm² A*)	$\bar{\alpha}$	$\sigma(V)$ (*meters per second*)
1	1×10^{-16}	0.9	153
2	2×10^{-16}	0.8	112
5	5×10^{-16}	0.6	90
10	10×10^{-16}	0.4	98
20	20×10^{-16}	0.2	144

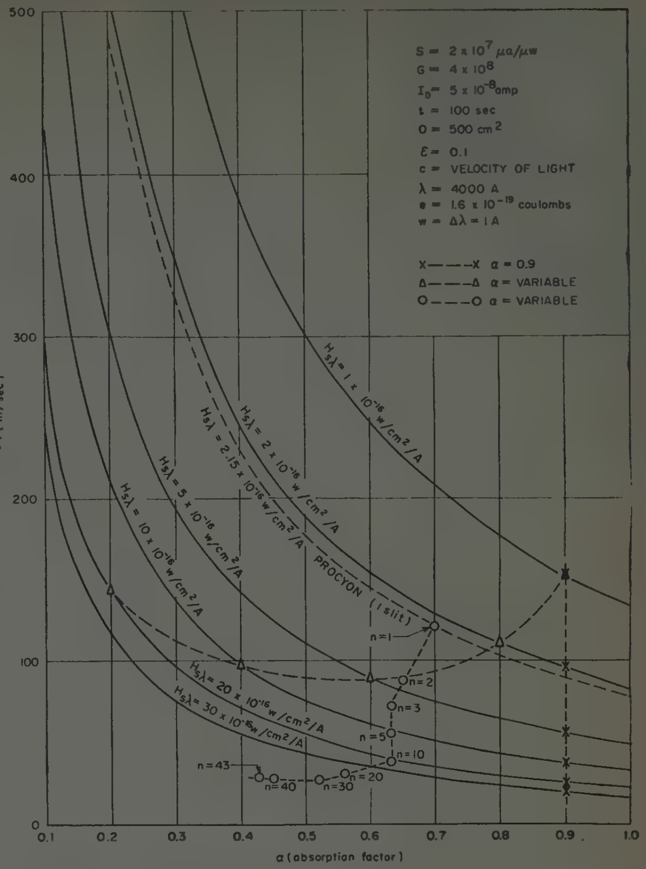

Fig. 7—Velocity errors vs absorption factor for various levels of stellar irradiance.

These points appear in Fig. 7 as △··· △. It is evident that for the values of α chosen for this illustration, the optimum number of lines is five or six. If more than this number are used, the velocity error increases.

This type of calculation was also performed using the spectrum of Procyon from 3990 A to 4020 A, part of which is reproduced in Fig. 3(a). It was assumed that the narrowest and most absorptive line would be chosen when only one slit is used. In this case, $\alpha=0.7$ and $\sigma(V)$ is 121 meters per second. This point is ○$n=1$ on the curve for $H_{s\lambda}=2.15\times10^{-16}$ which is the value in watt /cm²/A of the irradiance of Procyon at $\lambda=4000$ A. For the next most absorptive line, $\alpha=0.6$, reducing the average of α to 0.65. Similarly, the average $\bar{\alpha}$ was taken for the 43 lines in the spectral range under consideration and the velocity errors were calculated. These points appear in Fig. 7 as ○··· ○. In this case, little is to be gained from the use of more than ten lines.

THE HYDROGEN 21-CM LINE

The degree to which the hydrogen line may be employed in determining the velocity of a space vehicle is also dependent upon the intensity of the line, its sharpness, and the angular extent and location of the sources. Most of the lines found to date are of rather complex structure due to thermal motions and other turbulence within the clouds, and also to differences in radial velocity of the various clouds along the line of sight. In emission, the sharper lines range in width from 120 kc to 170 kc, corresponding to a velocity range of 24 km/sec to 34 km/sec. Quite often, the lines are much broader, and some widths are found ranging from 800 kc (160 km/sec) to 1 mc (200 km/sec). Absorption lines, where found, have characteristics much more suitable for Doppler measurement. Fig. 8 (see Hagen, *et al.*[7]) shows the line configuration obtained when viewing the Cassiopeia A source. The absorption lines exhibit widths at the half-power points of only 20 to 25 kc. The much increased sharpness of lines in absorption is attributed to the greater angular resolution obtained through the use of discrete sources. The angular size of the Cassiopeia A source is estimated at only 6 minutes of arc which would correspond to the beamwidth of a 500-foot antenna. This equivalent small beamwidth restricts the amount of hydrogen cloud viewed, so that smaller random velocity and Doppler shifts due to cloud motion are included.

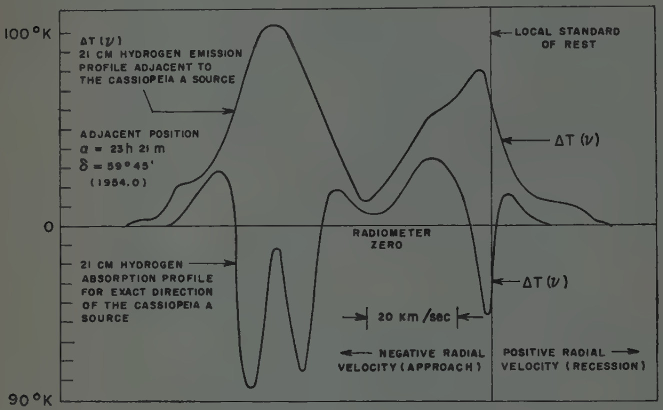

Fig. 8—Profile type of absorption measurement. (Two 21-cm hydrogen line profiles are shown, one for the direction of the Cassiopeia A source and one for a comparison position taken approximately 1° away.) (See Hagen, *et al.*[7])

The energy available from 21-cm sources is also small, necessitating the use of large antennas and highly stable radiometer-type equipment with long integration times, in order that the signal may be distinguished from the noise background. Brighter sources which have been discovered to date provide peak antenna temperatures in the neighborhood of 100°K. Most of the sources, however, provide energy levels much less than this. In order to provide a signal-to-noise ratio of at least 10.0, when the antenna temperature is 80°K the receiving equipment must be so designed that the equivalent noise temperature is below 8.0°K. Another factor entering into the receiver design is the band-pass. It has been determined that in order to provide good definition, the band-pass should be no larger than 5 kc. However, this is subject to some variation for our purposes, as we shall see later. Therefore, according to the definition of antenna temperature as Ta,

$$Ta = \frac{P}{k\Delta f} \tag{14}$$

where

P = power received by antenna,
k = Boltzmann's constant,
Δf = receiver band-pass.

If $\Delta f = 5$ kc, the input power corresponding to an antenna temperature of 80°K is only 5.5×10^{-18} watts, and the noise level for good measurement capability should not exceed approximately 10^{-19} watts.

The ordinary form of the radiometer equation postulated by Dicke[8] for the rms fluctuation in antenna temperature when looking at a radio source is

$$\Delta T = [(F - 1)T_0 + T_A]\pi/2 \sqrt{\frac{1}{B\tau}} \tag{15}$$

where

F = noise figure for signal and image bands,
B = predetection bandwidth,
τ = integration time,
T_0 = ambient temperature,
T_A = antenna temperature of radio source.

Noise figures of 8 db ($F = 6.3$) are readily realized, giving a receiver noise temperature $(F-1)T_0$ of 1550°K. Using a source antenna temperature of 80°K, and a 5-kc band-pass, the above equation shows that an integration time of 100 seconds results in a detectability ($S/N = 1$) of 3.6°K. Recent reports on the successful application of maser preamplifiers at 21 cm indicate that receiver noise temperatures at least as low as 50°K are realizable. In this case, a band-pass of 5 kc and an integration time of 100 seconds would permit a detectability of 0.3°K, showing an improvement of better than 10:1.

These figures on sensitivity can now be applied to the problem of determining the accuracy with which line shift, and hence velocity, can be measured. The method of detecting line position will be assumed to employ detection channels on either side of the line being measured, similar to the technique described for optical frequencies. Work has been done on optimum channel widths and separation to provide best sensitivity.[9] It will suffice to point out briefly that the bandwidth of the filters and the separation between them should be of the order of magnitude of the halfwidth of the line. Under these conditions, it can readily be shown that for small displacements Δf of the line, having a depth T expressed in units of antenna temperature, the power in one channel will increase approximately by an amount $kT\Delta f$, and the power in the second channel will decrease by a like amount, giving a total change in input power of $2kT\Delta f$. This will cause an equivalent temperature change at the input of the receiver of

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[7] J. P. Hagen, S. E. Lilley, and E. F. McClain, "Absorption of 21 cm radiation by interstellar hydrogen," *Astrophys. J.*, vol. 122, pp. 361–375; November, 1955. See p. 369.

[8] R. H. Dicke, "The measurement of thermal radiation at microwave frequencies," *Rev. Sci. Instr.*, vol. 17, pp. 268–275; July, 1946.

[9] P. M. Schultheiss, C. A. Wogrin, and F. Sweig, "Time frequency measurement of narrow band random signals in the presence of wide band noise," *J. Appl. Phys.*, vol. 25, pp. 1205–1036; August, 1954.

$$\Delta T = \frac{2kT\Delta f}{kB} = \frac{2T\Delta f}{B}, \qquad (16)$$

where B = the receiver band-pass.

Now it has been shown that with an integration time of 100 seconds, a detectability of 3.6°K is well within the capability of an ordinary radiometer, and with a maser preamplifier, detectabilities of at least 0.3°K are possible. The band-pass of the receiver should be adjusted as nearly as possible to correspond to the halfwidth of the line being measured. Referring once again to Fig. 8 it is found that this value for the Cassiopeia absorption line is approximately 4 km/sec which at the frequency of 1421 mc corresponds to 19 kc. T has the value of 80. Substituting these values in (16) and noting that increasing the receiver band-pass to 19 kc changes the detectability from 3.6°K to 1.8°K, we find an uncertainty in the measurement of line position of 45 meters per second (148 feet per second), and with a maser ($\Delta T = 0.15°$K), an uncertainty of only 3.7 meters per second (12.2 feet per second). In measuring velocity by means of Doppler shift, the difference between the shifted and the unshifted line must be determined so that the uncertainty in velocity measurement should be twice the above values or 90 meters per second (296 feet per second) and 7.4 meters per second (24.4 feet per second), respectively.

At the present time, the hydrogen line at 21 cm is still the only line that has been detected in interstellar space by radio techniques. A number of other lines have been predicted, however. Data thus far gathered indicate that the diatomic molecules, especially hydrides, are the important molecules of interstellar space. Radiation from these molecules has been predicted throughout the millimeter wave region, and far infrared. Unfortunately, the detection of milllmeter lines is very difficult compared with the 21-cm line because of equipment limitations, absorption in the terrestrial atmosphere, the reduced abundance of the molecules, the reduced-state temperature of the molecular energy levels defining the millimeter transitions and the lack of bright point sources in the sky at these frequencies. Improved equipment and satellites are certain to overcome the first two objections. Hence, the detection of these lines is not considered hopeless. At present, however, an appraisal of their value to navigational problems in space is impossible. It does appear fairly certain that, because of the rarity of these lines, the intensity of radiation received even in outer space at these wavelengths will be far less than that due to hydrogen, thereby raising considerably the performance level required of the equipment. Cosmic rays have also been examined, but have been found not to hold any immediate promise for space navigation, primarily because no discrete sources are available, and no discrete line spectra exist for the measurement of frequency shifts.

SOLAR RADIATION AND SOLAR DIAMETER

Solar radiation has been suggested as a means of determining the distance of a space vehicle from the sun. This information coupled with angle measurements on the sun and planets would permit the determination of vehicle position in space. Press has estimated that the total uncertainty in measuring the earth-sun distance with such a scheme would be 10,000 miles.[10] For velocity, our primary interest would be in the change in radiation power with time. Therefore, the absolute magnitude of solar radiation is unimportant. Constancy of radiation with time, on the other hand, is extremely important.

Power radiated from the sun and intercepted by a detector may be approximated from the black body law as

$$P = \frac{\sigma A_s A_D T^4}{\pi D^2}, \qquad (17)$$

where

σ = Stephan-Boltzmann constant,
A_s = area of sun,
A_D = area of detector,
T = temperature of the sun,
D = distance from the sun,

and the rate of change of this power with time is given by

$$\frac{dP}{dD} = \frac{2\sigma A_s A_D T^4}{\pi D^3}. \qquad (18)$$

Radial velocity with respect to the sun is therefore given by

$$V = \frac{\Delta P}{\Delta t} \times \frac{1}{\frac{dP}{dD}}. \qquad (19)$$

Table IV lists the radiation power and rate of change with distance for the neighborhoods of Earth, Venus, and Mars.

TABLE IV
(AREA OF DETECTOR = 1 CM²)

	Earth	Venus	Mars
P (watts	0.14	0.18	0.118
$\frac{dP}{dD}\left(\frac{\text{watts}}{\text{meter}}\right)$	1.74×10^{-12}	2.41×10^{-12}	1.58×10^{-12}

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[10] S. Press, "An application of solar radiation to space navigation," *Aero-Space Engrg.*, vol. 17, pp. 51–54; November, 1958.

The procedure in making measurements of radial velocity with respect to the sun would be that of taking distance measurements at two different times, and dividing the distance change by the time interval. Accordingly,

$$V = \frac{D_1 - D_2}{T_1 - T_2}. \qquad (20)$$

Assuming that errors exist in both the time and distance measurements, the resulting error in velocity is given by

$$\Delta V = \frac{\Delta D_1}{T_1 - T_2} + \frac{\Delta D_2}{T_1 - T_2} + \frac{(D_1 - D_2)\Delta T_1}{(T_1 - T_2)^2} + \frac{(D_1 - D_2)\Delta T_2}{(T_1 - T_2)^2},$$

and, assuming that $\Delta D_1 = \Delta D_2$, and $\Delta T_1 = \Delta T_2$,

$$\Delta V = \frac{2\Delta D}{T_1 - T_2} + \frac{2\Delta T(D_1 - D_2)}{(T_1 - T_2)^2}. \qquad (21)$$

It becomes apparent that the velocity error is dependent upon the time interval between readings. If the velocity may be considered as a slowly changing quantity so that readings may be taken over intervals which are large compared with the accuracy of measurement, the error may be controlled. For a time interval of 100 seconds, and a photocell detectivity of 10^{-6} watts (6×10^5 meters), the velocity uncertainty at the distance of the earth from the sun would become approximately

$$\Delta V = \frac{2 \times 6 \times 10^5}{100} = 12 \text{ km/sec (39,000 feet/second)}.$$

Turning now to the problem of measuring distance and velocity through angular measurements of the sun's diameter, the distance error for a given error in angular measurement is given by

$$\Delta D = \frac{d\Delta\alpha}{\alpha^2}, \qquad (22)$$

where

d = diameter of the sun,
α = angular diameter of the sun in radians.

Taking the earth-measured angular diameter of the sun as 0.0093 radian and the diameter as 13.9×10^8 meters, the error in the measurement of the earth-sun distance for each second error in angular measurement is

$$\Delta D = \frac{13.9 \times 10^8 \times \pi}{(0.0093)^2 \times 3600 \times 180}$$
$$= 7.8 \times 10^7 \text{ meters (48,500 miles)}.$$

Assuming an angular error as 0.1 second (which is realistic in view of present equipment),

$$\Delta D = 7.8 \times 10^6 \text{ meters}$$

and the error in determining radial velocity with a 100-second interval between readings would be given as before:

$$\Delta V = \frac{2\Delta D}{T_1 - T_2} = \frac{2 \times 7.8 \times 10^6}{100}$$
$$= 156 \text{ km/sec (510,000 feet/second)}.$$

CONCLUSIONS

Reviewing the results of the past study, a number of conclusions can be drawn. First of all, cosmic rays do not appear feasible as an aid to space navigation, primarily because no discrete sources of radiation are available and no discrete line spectra exist. Studies of other radiation in the visible and radio regions reveal that the primary difficulty is one of extremely low signal levels. To illustrate, the brightness temperatures of 21-cm sources range from 0° to 100°K. Viewing an 80°K source with a 50-foot antenna and a receiver band-pass of 5 kc, one would have a signal level of only 5.5×10^{-18} watts. At optical wavelengths, the power available from the brightest stars in the region of 5000 A is approximately 1×10^{-12} watts/A if viewed through a 10-inch telescope. The sun, of course, is an exception. When viewed under the same conditions, a quite adequate power level of 9.5×10^{-3} watts/A is obtained.

Except for the sun, then, it is obvious that in measuring Doppler shifts with such low-level signals, special detection equipment with high sensitivity and long time constant is required in order to distinguish the signal from the noise. If the assumption is made that the detection limit occurs when the signal is equal to the noise, the capabilities of various systems can be compared as in Table V where the noise level is expressed in

TABLE V
SYSTEM COMPARISON

Source	System	Integration Time	Noise Level
Optical Doppler			
Sun	Babcock Magnetograph	100 seconds	3 feet/second
Sirius	10-inch Telescope Detection System Similar to Magnetograph	100 seconds	60 feet/second
Radio Doppler (21 cm)			
Cassiopeia A	Radiometer 50-foot Antenna	100 seconds	148 feet/second
Cassiopeia A	50-foot Antenna with Maser	100 seconds	12.2 feet/second
Other			
Sun	Total Radiation	100 seconds	39,000 feet/second
Sun	Diameter Measurement	100 seconds	510,000 feet/second

terms of equivalent velocity in feet per second. This noise level varies inversely as the square root of the output integration time for the radio and optical systems. Consequently 100 seconds was chosen as a convenient value upon which to base our comparison. In instances where longer times can be tolerated, an equivalent improvement in detectability may be expected up to the point where equipment instabilities become a problem. It should be pointed out that where velocity is derived from measurements of total solar radiation or solar diameter, the error varies inversely as the time interval between measurements. So-called "integration time" has therefore a special connotation in these cases.

Taking into account the number of available sources, the energy available, and the complexity and size of equipment, it appears that natural radiation in the optical range of frequency holds the greatest promise for early successful application to space navigation. Excellent detectors are available in this range, but low signal level still persists as a major problem.

The Optimization of Astronautical Vehicle Detection Systems Through the Application of Search Theory*

NORMAN S. POTTER†, SENIOR MEMBER, IRE

Summary—**A generalized theory of detection system performance is developed and applied to the analysis of collision warning and the optimal allocation of search effort for astronautical vehicles. The kinematic basis of the relative frequency of intercepts with randomly moving particles is presented. The pronounced variation in warning system effectiveness if the possible contact space is uniformly scanned is displayed by investigating the functional behavior of the weighted mean acquisition range and cumulative probability of detection by some minimal detection barrier in the hypervelocity closure rate, low signal amplitude operational environment. Fixing the search system frame time, it is shown that the probability of acquiring a closing particle by some critical range determined by the system response delay may be made independent of the bearing angle of the relative approach path by choosing search system dwell times that are proportional to the second or fourth power of the mean anticipated closure rate, the exponent being dependent upon whether the sensor is active or passive.**

The rational choice of the frame time is investigated in terms of the information rate out of the sensor and the requirements of the decision-making apparatus. The latter is assumed to be an unsaturated, bandwidth and memory limited data processing subsystem. The probability of track retention and the variance of the best estimate of the contact-bearing angle are related to the cumulative probability of detection. It is shown that the required system reaction times are in conflict with the optimal detection system information rates.

The very important case of contacts between astronauts and other objects independently moving on trajectories defined by a generalized central force field is separately considered. It is shown that while a true collision contact is only possible in trivial cases, as viewed by the astronautical vehicle the relative terminal approach trajectory of the particle is essentially confined to a surface defined by reasonable navigational system outputs. The confinement of the search to this restricted region of contacts makes the problem of detection in such an unfavorable environment far more amenable to solution than is possible with omnidirectional coverage.

I. INTRODUCTION

THE obvious stringent weight and space restrictions which obtain in the design development of conventional airborne detection systems are necessarily accentuated if the equipments are intended for astronautical vehicles. Unfortunately, while the means of solution to problems such as those of anticollision warning decline with the possibility of implementing effective sensors, the situation with regard to astronauts is further aggravated by the hypervelocity closure rates and minuscule size of meteors that will frequently be the subjects of detection. Since sensor improvements offer little in the way of adequate enhancement of the signal levels themselves, alternate solutions in terms of the extraction of more information from such signals as may be present, or an increase in the efficiency of the

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detection process through optimal allocation of the search effort, must be sought. This paper is concerned with the latter.

To this end, statistical models which are descriptive of most cases of interest, or at least those that are amenable to general analysis, are investigated. As an approximation to the kinematic environment created by sporadic meteors, rectilinear randomly-oriented motion of observer and targets is assumed. It is reasonable to expend the greatest search over those regions of the relative observer space that are the most likely sources of contacts. The question as to how this is best accomplished will be examined at length with the aid of the first statistical model, and an optimal scan program which can be computer-ordered through operation upon navigation system inputs will be derived.

The preceding model is of utility in discussing search during near-terrestrial powered flight and, indeed, appears to provide a meaningful description of the probability density of contacts for conventional vehicles. Representative problems for the latter, to which much of what is to follow is applicable, are the provision of anticollision warning, or the determination of the random target search to be performed by an interceptor in the absence of adequate ground control. However, for reasons of fuel economy, the overwhelmingly greater portion of any space flight must be unpowered and would tend to involve trajectories that are largely described by the kinematics of motion in an isolated central force field. Accordingly, the purely random distribution is later discarded and the nature of contacts between objects that are randomly distributed in space, but whose motion reflects the bias introduced by a communal gravitational field, is investigated.

The issue of anticollision warning has been touched upon above. The density of objects of other than natural origin is now vanishingly small, and undoubtedly will remain so for any foreseeable future. Beyond this, there is an understandable diversity of opinion regarding the frequency and magnitude of the danger of contacts with meteors. Tabulations of earth impacts markedly favor the extremely small particles and suggest, in highly qualitative, generalized terms, that impacts of astronautical vehicles with particles of sufficient size to be catastrophic should be comparatively rare. However, the extent of the commitment of the national industrial potential and human resources to any space exploration enterprise is such as to motivate against the admission of elements of chance. In brief, no rationale is needed at this time for the theoretic consideration of the warning problem in a study devoid of commentary upon specific sensor equipments, or the various courses of action that suggest themselves should such warning be accomplished. Indeed, the feasibility, or even desirability, of responses such as an evasive perturbation of the trajectory of the astronaut must remain unexplored.

II. The Statistics of Particle Distribution and Detector Performance

A. The Density of Contacts as a Function of Bearing Angle

The distances involved are so vast that, over the more modest detection ranges attainable by any sensor that can be envisioned for such application, and the moderate sampling times that would obtain, observable motions will be effectively linear. Further, at least for particles originating in our solar system, their velocities will tend to span some reasonably narrow spectrum of values in a given region of space. Accordingly, it is not unreasonable to construct the required statistical model upon the hypothesis of rectilinear motion of all objects under consideration, and to assign to the particles some mean representative velocity. The latter is not a necessary restriction and, should it later prove possible to assign some probability density function to the velocities that may obtain, the results could be appropriately modified to yield a joint density function of the velocity and the bearing angle β. The phrase "bearing angle" is used in this paper to indicate the angle described by the longitudinal axis of the vehicle and the radial line from the observer to the point in space in question, and measured aft from the positive sense of the velocity vector. In particular, it is not intended to infer thereby that planar motions alone are under consideration.

The mean number of collision contacts to be made along a given bearing line, or rather in a solid angular cone of apex angle $\delta\beta$, is readily shown to be proportional to the closure rate at β. To demonstrate this, we consider a sphere of contacts of arbitrary radius ρ, and note that no particle of a given heading angle can enter unless its relative velocity vector $\boldsymbol{V}$ is so oriented as to point in toward the sphere. That is, the angle between the negative of $\boldsymbol{V}$ and the unit radial vector $\boldsymbol{n}$ must be acute. To a first-order differential approximation, all contacts made in some unit time interval must lie within the relative space volume $(-\boldsymbol{V}\cdot\boldsymbol{n})\rho^2\delta\beta$ which is swept by the moving observer. The required mean number is found by multiplying the preceding expression by the product of the space density of particles and the probability density of the heading angle in question, both of which are constant by virtue of the assumption of randomness. If, in particular, the particle is on a collision course, the normal and reversed relative velocity vectors are coincident and the scalar product is reduced to $|\boldsymbol{V}|$, which was to be shown. It should be noted that the resultant expression is functionally dependent upon β only to the extent that the closure rate is dependent. Koopman [1] extensively discusses the above for planar motion and develops the kinematic basis of detection more fully than is required here.

Certain nonquantitative observations can be made at this time to clarify the intent and structure of the subsequent discussion. It will be shown presently that, all else being equal, the cumulative probability of detection by some minimal admissible range declines very markedly with closure rate. Accordingly, the relative velocity, and therefore the distribution of the search effort, figure prominently. From the above, performance is poorest where it is needed most by virtue of the distribution of the mean density of contacts, unless the allocation of detector dwell is optimized in some sense. To this end, a statistical model of detector performance will now be developed.

B. The Analytical Model of an Idealized Detector and System Optimization

In the brief investigation which follows, no appraisal will be made of such considerations as the significance of the total information content of a sequence of signals. Techniques of noise discrimination through kinematic analysis of the implied motion of a signal source over some sampling interval are representative of such refinements. However, highly sophisticated data processing equipments involving extensive system memory and computer bandwidth would appear to be largely precluded from application in astronautical vehicles. Accordingly, the detection criteria assumed here consist of, at most, a redundancy counter or its equivalent.

Empirically derived data and theoretic analyses suggest that even in the absence of the scintillation induced by a physically dense propagating medium, the input signal power is Rayleigh distributed. That is, the probability density of the scan-to-scan fluctuation is $\overline{S}^{-1} \exp(-S/\overline{S})$, S and $\overline{S}$ denoting, respectively, the instantaneous and mean received signal powers. The dwell time of the scanner will necessarily be so brief in this instance that a high order of signal correlation over that interval may be assumed, in contrast with the individual interrogations which may be regarded as statistically independent trials. Integrating the density function from some threshold S_m to the point at infinity, the single-glimpse probability, exclusive of later information losses, is found to be of the form $\exp(-S_m/\overline{S})$.

The preceding material, as well as the analysis set forth in this section, has been discussed far more extensively in Potter [2], which also investigates the issue of the performance degradation caused by data-handling subsystem limitations that lead to information loss. Though not discussed here, unavoidable restrictions upon computer sophistication must degrade the system performance.

Let the cumulative probability of detection P at range R be required. Writing the single glimpse probability at the λth interrogation as $\psi \equiv \psi[R+\delta+(n-\lambda)Vt]$, where V is the relative range rate, t is the search period, the time interval between successive samples of a given subsector of the relative space, and δ is a small range increment which, physically interpreted, reflects the fact that the last scan may, with equiprobility, occur anywhere within the interval

$$(R, Vt + R), \qquad P = 1 - \prod_{\lambda=1}^{n} (1 - \psi_\lambda),$$

or

$$\ln(1 - P) = \sum_{\lambda=1}^{n} \ln(1 - \psi_\lambda). \tag{1}$$

The introduction of δ will now be seen to be a device to effect the transition from a discrete to a continuous functional representation since the expected value of $\ln(1-P)$ is its mean with respect to δ taken over the half-open interval over which δ is defined. It may be shown through application of the mean value theorem [2] that if, as is the case in this instance, the sampling rate is high the asymptotic relationship $\overline{\ln(1-P)} \approx \ln(1-\overline{P})$ obtains. As used here, the bar superscript designates a weighted mean value. Assuming, then, the validity of the preceding inferred asymptotic behavior,

$$\begin{aligned}\ln(1 - \overline{P}) &= \frac{1}{Vt} \sum_{\lambda=1}^{n} \int_0^{Vt} \ln\{1 - \psi[R + \delta + (n - \lambda)Vt)]\} \partial\delta \\ &= \frac{1}{Vt} \sum_{\lambda=1}^{n} \int_{R+(n-\lambda)Vt}^{R+(n-\lambda+1)Vt} \ln[1 - \psi(\rho)] d\rho \end{aligned} \tag{2}$$

or, discarding the bar notation and accomplishing the indicated summation,

$$\begin{aligned}P &= 1 - \exp\left[\frac{1}{Vt} \int_R^{R+nVt} \ln(1 - \psi(\rho)) d\rho\right] \\ &= 1 - \exp\left[\frac{1}{Vt}\left(\int_R^\infty - \int_{R+nVt}^\infty\right) \ln(1 - \psi(\rho)) d\rho\right] \\ &= 1 - (1 - P_R)/(1 - P_{R+nVt}), \end{aligned} \tag{3}$$

where P_R, P_{R+nVt} denote cumulative detection probabilities for search from the point at infinity to the ranges indicated in the subscripts. Since nVt, the relative range traversed during the detection process, is large, P_{R+nVt} is necessarily small with respect to P and P_R, so that one must have $P \approx P_R$.

Adopting the latter approximation, one may now conveniently particularize the resultant general expression to the case of Rayleigh signal power densities. Writing $S_m/\overline{S} = \alpha\rho^\nu$, where ν is 2 or 4 according to whether the sensor is passive or active, α is a parameter that is functionally dependent upon detector and particle characteristics, and ρ is their instantaneous relative range,

$$
\begin{aligned}
P &= 1 - \exp\left[\frac{1}{Vt}\int_R^\infty \ln(1 - e^{-\alpha\rho^\nu})d\rho\right] \\
&= 1 - \exp\left[-\frac{R}{Vt}\left\{\frac{1}{R\alpha^{1/\nu}}\int_{R\alpha^{1/\nu}}^\infty \ln(1 - e^{-\rho^\nu})^{-1}d\rho\right\}\right] \\
&= 1 - \exp\left[-\frac{R}{Vt}N(x)\right], \qquad (4)
\end{aligned}
$$

where $x = R\alpha^{1/\nu}$, and

$$
N(x) = \frac{1}{x}\int_x^\infty \ln(1 - e^{-\rho^\nu})^{-1}d\rho. \qquad (5)
$$

Eq. (4) may be cast into a somewhat more convenient form which relates the single glimpse and cumulative detection probabilities at some specified range to one another by noting that, for Rayleigh signals, if ψ_R denotes the single glimpse probability at range R, at any other range ρ

$$
\psi(\rho) = \psi_R{}^{(\rho/R)^\nu} = \exp\left[-\rho^\nu(\ln\psi_R{}^{-1})/R^\nu\right].
$$

Accordingly,

$$
\alpha = (\ln\psi_R{}^{-1})/R^\nu,
$$

and

$$
P = 1 - \exp\left[-\frac{R}{Vt}N(\ln^{1/\nu}\psi_R{}^{-1})\right]. \qquad (6)
$$

Certain observations regarding the choice of scan rates in automatic tracking systems may now be made. Detection is generally followed by at least a brief interval of tracking which may even be integral with the detection process if the decision as to the reality of a contact is predicated upon whether or not its apparent trajectory is erratic. At least in anticollision systems, trajectory analysis is necessarily inherent to the detection process, since contacts of interest are characterized by consistent relative closure. However, the optimization of the sampling rate to yield the greatest attainable cumulative probability of detection in the classical sense, the assurance of a recognizable signal by some specified relative range, will be seen to be at wide variance with the sampling rates indicated by tracking considerations alone. The latter will now be discussed, while optimization of the cumulative detection probability will be deferred to a subsequent portion of this paper.

The admissible prediction time, the interval over which the implied track of an object is projected between successive scans or attempts at correlation, is necessarily limited in tracking systems of the type that would be appropriate for inclusion in an astronautical vehicle. The sophistication of the scan-to-scan correlation process and, in turn, computer bandwidth, grows with the prediction interval, since the region of uncertainty or envelope of potential positions of the contact is a monotone increasing function of the time between successive signals. Accordingly, the expected number of detections relating to a particular contact per unit period of time provides a measure of performance in the large class of systems in which the likelihood of retention of track and the quality of the data processing subsystem output are critically dependent upon the density of the sequence of available samples.

In what is to follow, T denotes the time interval of interest, while the performance of the detector at the given range is defined by the single glimpse probability ψ_R, attained with a scan rate such that successive interrogations occur at intervals t. If, through variation of the scan rate, this latter quantity assumes the value t', then, if the mean signal to root-mean-square noise voltage ratio of the integrator output is taken as proportional to $\sqrt{t'}$, n, the expected number of successful interrogations of the contact may be approximated by $T/t'\,\psi^{t/t'}$. As is apparent, the validity of the preceding expression is contingent upon an interrogation rate that is sufficiently high to insure that the change in the observer-contact relative geometry is inappreciable over the total sampling time.

The first and second derivatives of $n(t')$ with respect to t' may be exhibited in the form

$$
\begin{aligned}
\frac{\partial n}{\partial t'} &= -\frac{T}{t'^2}\psi^{t/t'}\ln(\psi^{t/t'}e), \\
\frac{\partial^2 n}{\partial t'^2} &= -\frac{1}{t'}\frac{\partial n}{\partial t'}\left(2 + \frac{t}{t'}\ln(\psi e)\right) - \frac{Tt}{t'^4}\psi^{t/t'}; \qquad (7)
\end{aligned}
$$

$n'(t')$ vanishes at the origin and the point at infinity, and it, as well as its first derivative, is continuous elsewhere. Therefore, the extremals are the positive, finite zeros of the derivative. The latter vanishes if, and only if, the logarithmic term does, indicating that, all else fixed, the dwell time during an individual interrogation should be so chosen as to yield a single glimpse probability of e^{-1} at the desired range, if the mean time between successive contacts is to be minimized. That the choice is the unique optimum in this sense follows immediately upon noting that the second derivative is negative definite when the first derivative vanishes.

Before turning briefly to the analysis of the quality of the output data as a function of the information rate provided by the sensor, it should be pointed out that other reasonable optimization criteria may be hypothesized. For example, instead of seeking to maximize the number of successful interrogations of the contact obtained over some time interval, as has been done above, the sampling rate may be so chosen as to maximize $\bar{P}$, the probability that at least one successful interrogation results. That is, because of stringent bandwidth and storage restrictions, the data processing logic may be so arranged as to discard the trajectory if, on account of a continuous succession of fruitless interrogations, the prediction time, measured from the last contact, exceeds some predetermined figure.

It is apparent that, subject to the same assumptions and notation used above,

$$\tilde{P} = 1 - (1 - \psi_R{}^{t/t'})^{T/t'}, \tag{8}$$

partial differentiation of which with respect to t' shows that the extremals are the zeros of

$$y = (1 - \psi_R{}^{t/t'}) \ln (1 - \psi_R{}^{t/t'}) - \psi_R{}^{t/t'} \ln \psi_R{}^{t/t'}. \tag{9}$$

From inspection, they are determined by allowing the resultant single glimpse probability to assume the values 0, $\frac{1}{2}$, and 1. The first and last are physically meaningless, since the indicated sampling intervals are, respectively, infinitesimal and infinite. Indeed, consideration of the limiting behavior of (8) shows that they are minima so that, by virtue of the continuity of $\tilde{P}$, it must follow that system performance is uniquely optimized in this sense by choosing the signal integration times in such a manner that, all else fixed, it is equiprobable that an individual interrogation succeed or fail at the desired relative range.

The conflict among the optima is apparent since the ratio of the maximum information rate sampling interval to that derived immediately above is ln 2 for both passive and active search. Further, if P_0, t_0 and $\tilde{P}$, $\tilde{t}$ denote the cumulative detection probabilities and sampling intervals associated respectively with the former and the latter choices of optimization criteria, (6) yields

$$P_0 = 1 - \exp\left[-\frac{R}{Vt_0}\cdot N(1)\right],$$

$$\tilde{P} = 1 - \exp\left[-\frac{R}{V\tilde{t}} N(\ln^{1/\nu} 2)\right], \tag{10}$$

from which, eliminating R/V,

$$P_0 = 1 - (1 - \tilde{P})^{(1/\ln 2)N(1)/N\ (\ln^{1/\nu} 2)}. \tag{11}$$

Numerical analysis of the function $N(x)$ reveals that the exponent $N(1)/N(\ln^{1/\nu} 2)$ ln 2 is 0.67 or 0.83, according to whether the search conducted is active or passive. However, in both instances, it then follows that $\tilde{P}$ always exceeds P_0.

The specific nature and quality of the data to be derived by the sensor-computer may vary considerably from estimation of the bearing angle alone to comparatively sophisticated analyses of the implied trajectory of the contact. If the errors in the determination of any appropriate, related quantity are randomly distributed in each observation and a best estimate is to be formed upon the basis of some sufficiently lengthy sequence of interrogations, a reasonable figure of merit for the appraisal of system performance is the influence upon the variance of the failure of the sensor or the data processing subsystem to obtain or achieve the correlation of the necessary observations. Alternately stated, as the number of available observations that relate to a given contact decline, the variance of the resultant best estimate of the desired quantity must increase.

The further statistical study of the choice of the rate at which the potential contact space is to be interrogated requires a sampling theorem [3] which states that if the variance and the probability of obtaining each of a series of n observations is fixed, which is in accord with the assumption that the observer-contact relative geometry is essentially invariant during the sampling interval, the ratio ρ of $\sigma_p{}^2$, the anticipated variance of the best estimate based upon a sequence of trials whose length is statistical, to σ^2, the idealized variance based upon estimates in which all interrogations were successful is, for n large, inversely proportional to the probability p of obtaining an individual observation. If the data processing subsystem is assumed not to degrade the sensor-provided information, p is precisely the single glimpse probability of detection ψ. In a sense, the preceding states that, as the idealized number of samples increases, the expected variance of the resultant best estimate is inversely proportional to the expected number of samples available for analysis. Demonstration is required despite the obvious appeal to instinct, since it is only infrequently that the expected value of a function of a statistically distributed variable is equal to the function of the expected value of the variable.

In this instance the function is $1/x$, where the argument assumes positive integral values, λ, the probability of λ being

$$\psi^\lambda(1 - \psi)^{n-\lambda}\binom{n}{\lambda},$$

since its distribution is Bernoullian.

Writing σ^2 in the form k/n, where k is some unspecified function of the particle and the sensor-computer system, and forming the weighted mean of expressions of the form k/λ, $1 \leq \lambda \leq n$,

$$\rho = \frac{\sigma_p{}^2}{\sigma^2} = \frac{1}{1 - (1 - \psi)^n} \cdot \sum_{\lambda=1}^{n} \frac{k}{\lambda}\binom{n}{\lambda}\psi^\lambda(1 - \psi)^{n-\lambda} \Big/ \left(\frac{k}{n}\right). \tag{12}$$

Since ψ must be bounded away from zero if the problem is to be physically meaningful, the cumulative probability of obtaining at least one observation, appearing in the denominator of the right side of (12), approaches unity in the limit. Replacing n/λ by the exponential integral appearing in the summation below, it suffices to consider

$$\rho \simeq \sum_{\lambda=1}^{n}\binom{n}{\lambda}\psi^\lambda(1 - \psi)^{n-\lambda} \int_0^\infty e^{-t\lambda/n}dt. \tag{13}$$

Upon inversion of the order of summation, (13) assumes the form

$$\rho \simeq \int_0^\infty \left\{ \sum_{\lambda=1}^{n} (\psi e^{-t/n})^\lambda (1 - \psi)^{n-\lambda}\binom{n}{\lambda}\right\} dt$$

$$= \int_0^\infty \{[1 - (1 - e^{-t/n})\psi]^n - [1 - \psi]^n\} dt. \tag{14}$$

Neglecting for the moment the issue of the admissability of the inversion of the order of integration and the limiting process, since $[1-\psi]^n$ vanishes in the limit, only the behavior of

$$\phi(n) = [1 - (1 - e^{-t/n})\psi]^n \tag{15}$$

for n large is of interest. Taking the logarithm of both sides of (14), application of l'Hospital's rule yields

$$\ln \phi = \frac{\ln [1 - (1 - e^{-t/n})\psi]}{1/n}$$

$$\simeq \frac{-n^2\psi\left(\dfrac{t}{n^2}e^{-t/n}\right)}{1 - (1 - e^{-t/n})\psi} \simeq -\psi t. \tag{16}$$

Accordingly, the integrand goes to $e^{-\psi t}$ whence, by inference, $\rho \simeq 1/\psi$, which was to be shown. However, for all n the integrand is continuous and approaches its limit uniformly over any finite subinterval of the region of integration. Further, for all α, $0<\alpha<\psi$, $e^{-\alpha t}$ is integrable and dominates $e^{-\psi t}$ for positive t and, in turn, the integrand. These conditions are sufficient for the inversion of the limit to be admissible, thereby establishing the result.

In an idealized system, the signal-to-noise voltage ratio and, in turn, the standard deviation of the individual estimate of a directly observed quantity, such as the bearing angle of the contact, may be taken as inversely proportional to the square root of the integration time. On the other hand, the standard deviation of the best estimate varies inversely as the square root of the number of observations, where the latter is directly proportional to the sampling rate if, as has been hypothesized, the time interval over which the sequence of interrogations must be made is specified. It follows, then, that the variance of the best estimate obtainable, if all the interrogations are successful, is independent of the sampling rate. Accordingly, the statistical theorem derived above states that the dependence of the anticipated variance upon the sampling rate is determined only by the functional dependence of the single glimpse probability upon that latter quantity. In particular, the variance of the best estimate increases with the sampling rate. By way of an example, the variance in a maximum information rate system will be greater by a factor of $e/2$ than that in which the probability of at least one observation during the stipulated time interval is optimized.

It is apparent from the preceding discussion that the hypervelocity closure rates and short reaction time requirements, which necessitate high sampling rates, substantially influence detection system performance beyond the more obvious purely geometric degradation that will be investigated next. Certainly the choice of an appropriate sampling rate is seen not to be an elementary consideration in this instance and, from the preceding, it should be as small as is consistent with the over-all response times of the detection system and the kinematic behavior of the astronautical vehicle.

III. SOME IMPLICATIONS OF THE PRECEDING

A. The Dependence of Detection Performance Upon the Closure Rate

The influence of the unfavorable operational environment defined by hypervelocities and low signal amplitudes may be displayed in several ways. Qualitative physical reasoning as well as the preceding quantitative analysis indicate that the mean acquisition range $\overline{R}$ will decline as the situation is further degraded. Utilizing the continuous representation of the detection process developed above, the probability density of acquisitions is the two-branch function $-(dP/dR)/P(0)$ for $R \geq 0$ and the null function for R negative. $\overline{R}$ then assumes the form

$$\overline{R} = -\frac{1}{P(0)}\int_0^\infty \left(R\frac{dP}{dR}\right)dR$$

$$= \frac{1}{P(0)}\left\{\int_0^\infty P(R)dR - \lim_{R\to\infty} RP(R)\right\}, \tag{17}$$

the right-hand side of which results from integration by parts. $P(R)$ is a bounded function so that $RP(R)$ vanishes at the lower limit. Further, $P(R)$ goes to zero for R large and the logarithmic function in the integrand in $P(R)$ may be replaced by its exponential part. L'Hospital's theorem may then be applied to

$$\lim_{R\to\infty}\left\{1 - \exp\left[\frac{1}{Vt}\int_R^\infty \ln(1 - e^{-\alpha\rho^\nu})d\rho\right]\right\} \bigg/ \frac{1}{R}$$

$$\approx \lim_{R\to\infty}\left\{1 - \exp\left[-\frac{1}{Vt}\int_R^\infty e^{-\alpha\rho^\nu}d\rho\right]\right\} \bigg/ \frac{1}{R}\cdot \tag{18}$$

The differentiation process yields $R^2e^{-\alpha R^\nu}/Vt$, which goes to zero so that the weighted mean detection range is merely the ratio of the integral of the cumulative probability function over the entire positive half axis and its value at the origin.

Let $\overline{R}(v)$ and $\overline{R}(V)$ represent mean ranges derived under conditions of closure rates v and V, respectively. If the ratio of the velocities v/V assumes some constant value k, but each becomes indefinitely large, the nature of the dependence of $\overline{R}$ upon the relative velocity may be investigated by examining the behavior in the limit of the ratio,

$$\frac{R(v)}{\overline{R}(V)} = \left\{\frac{1 - e^{(1/Vt)F(0)}}{1 - e^{(1/kVt)F(0)}}\right\}$$

$$\cdot\left\{\frac{\displaystyle\int_0^\infty [1 - e^{(1/kVt)F(R)}]dR}{\displaystyle\int_0^\infty [1 - e^{(1/Vt)F(R)}]dR}\right\}, \tag{19}$$

where $F(R)$ denotes the logarithmic integral which appears in $P(R)$.

As V becomes sufficiently large, the exponentials in (19) may be replaced by the first two terms of their Maclaurin developments, from which the product of the bracketed expressions is seen to be unity. Accordingly, the sensitivity of the weighted mean range to moderate changes in closure rate, such as would be introduced by variation of the bearing angles under consideration, declines at hypervelocities.

In general, sophistication of the analysis of detection system performance is readily shown to be of declining utility with small signal amplitudes in that the weighted mean range ceases to be a complex function of the relative motion, and increasingly conforms to the elementary range equations. That is, considering $\overline{R}$ as a function of the signal power $\overline{S}$, a change to $k\overline{S}$ alters the initial mean acquisition range by a factor of $k^{1/\nu}$. This follows immediately from an argument similar to that used above. Writing

$$\frac{\overline{R}(k\overline{S})}{\overline{R}(\overline{S})} = \left\{ \frac{1 - \exp\left[\frac{1}{Vt}\int_0^\infty \ln\left(1 - e^{-a\rho^\nu/\overline{S}}\right)d\rho\right]}{1 - \exp\left[\frac{1}{Vt}\int^\infty \ln\left(1 - e^{-a\rho^\nu/k\overline{S}}\right)d\rho\right]} \right\}$$

$$\cdot \left\{ \frac{\int_0^\infty \left[1 - \exp\left[-\left(\frac{R}{Vt} N\left(Ra^{1/\nu}/(k\overline{S})^{1/\nu}\right)\right)\right]\right] dR}{\int_0^\infty \left[1 - \exp\left[-\left(\frac{R}{Vt} N\left(Ra^{1/\nu}/\overline{S}^{1/\nu}\right)\right)\right]\right] dR} \right\}, \quad (20)$$

which, following a change of variable of the form $R/c=\phi$ in the above integrals, may be written

$$\frac{\overline{R}(k\overline{S})}{\overline{R}(\overline{S})} = k^{1/\nu} \left\{ \frac{1 - \exp\left[\frac{\overline{S}^{1/\nu}}{Vt}\int_0^\infty \ln\left(1 - e^{-a\phi^\nu}\right)d\phi\right]}{1 - \exp\left[\frac{(k\overline{S})^{1/\nu}}{Vt}\int_0^\infty \ln\left(1 - e^{-a\phi^\nu}\right)d\phi\right]} \right\} \cdot \left\{ \frac{\int_0^\infty \left[1 - \exp \frac{(k\overline{S})^{1/\nu}}{Vt}\int_\phi^\infty \ln\left(1 - e^{-at^\nu}\right)dt\right]d\phi}{\int_0^\infty \left[1 - \exp \frac{\overline{S}^{1/\nu}}{Vt}\int_\phi^\infty \ln\left(1 - e^{-at^\nu}\right)dt\right]d\phi} \right\}. \quad (21)$$

As $\overline{S}$ becomes sufficiently small, the exponentials in (21) may be replaced by the first two terms of their Maclaurin development, leading directly to the required result. It should be noted that the conclusion is not contingent upon the assumption of a Rayleigh model, since the reasoning would proceed in an identical manner for all $\psi(\overline{S}/r^\nu)$, where r is the instantaneous relative range.

Unfortunately, while the preceding observations on the behavior of the weighted mean range present an interesting aspect of the detection process, it is very limited in scope. A more meaningful measure of system performance is the influence of the environment upon the cumulative probability, since survival is dependent upon detection outside some range barrier that is largely dictated by response delays introduced by the decision-making apparatus. Before introducing this latter consideration, some review of fundamental system behavior with respect to $P(V)$ is advantageous.

Fixing the range for the moment and examining the limiting behavior of $P(v)/P(V)$, as had been done above in the case of the weighted mean acquisition range, from (4),

$$\frac{P(v)}{P(V)} = \left\{1 - e^{-(R/kVt)N(R\alpha^{1/\nu})}\right\} / \left\{1 - e^{-(R/VE)N(R\alpha^{1/\nu})}\right\}, \quad (22)$$

which, for V large, is asymptotic to k^{-1}. Accordingly, at hypervelocities, the cumulative detection probability displays an approximate inverse dependence upon the closure rate. This is in marked conflict with the likelihood of potential collision contacts, which is proportional to the relative velocity. The preceding observation obtains whether the system performance degradation is caused by extreme closure rates or reduced signal amplitudes. This is readily established by recalling that α varies inversely as the mean signal power density, and by noting that $N(x)$ vanishes in the limit for x large. The result then follows by application of l'Hospital's theorem or by considering the magnitude of the terms in the Maclaurin development of the exponentials in (22) in powers of their exponents.

In actuality, of course, the degradation caused by even moderate increases in relative velocity is more substantial than has been indicated above, since, as has been mentioned earlier, at a given bearing the required detection range should be very nearly proportional to the corresponding closure rate. It is readily shown that the velocity dependence is then of a much higher order than that displayed above by again investigating the limiting behavior of $P(v)/P(V)$. In this instance,

$$\frac{P(v)}{P(V)} = \frac{1 - \exp\left[\frac{1}{kVt}\int_{kV\mathfrak{J}}^\infty \ln\left(1 - e^{-\alpha\rho^\nu}\right)d\rho\right]}{1 - \exp\left[\frac{1}{Vt}\int_{V\mathfrak{J}}^\infty \ln\left(1 - e^{-\alpha\rho^\nu}\right)d\rho\right]}$$

$$\sim \frac{1}{k} \frac{\int_{kV\mathfrak{J}}^\infty \ln\left(1 - e^{-\alpha\rho^\nu}\right)d\rho ,}{\int_{V\mathfrak{J}}^\infty \ln\left(1 - e^{-\alpha\rho^\nu}\right)d\rho}, \quad (23)$$

where $\mathfrak{J}$ is the over-all time constant of system response to the warning, and the indicated asymptotic behavior follows immediately from the observation that the exponents that appear are very rapidly decreasing functions of V. Differentiation of numerator and denominator and replacement of the resultant logarithmic expressions by the first term of their Maclaurin development leads to

$$\frac{P(v)}{P(V)} \sim {}^{\alpha(V\mathfrak{J})^{\nu}(1-k^{\nu})}. \tag{24}$$

Unlike the earlier fixed range example, the above approaches a limit for large V if, and only if, k is greater than unity. Physically, this suggests that even modest increases in closure rate will produce a very marked decay in the system performance, as is evidenced by the now exponential dependence upon that variable.

It is of interest to obtain a somewhat more quantitative insight into the behavior of $P(V\mathfrak{J})$. Assuming the preceding to be known, one might then inquire as to the cumulative detection probability p corresponding to some new closure rate v. From (6),

$$\psi_R = \exp\left\{-\left[N^{-1}\left(\frac{t}{\mathfrak{J}} \ln (1-P)^{-1}\right)\right]^{\nu}\right\}. \tag{25}$$

But the single glimpse detection probability at the new required detection range is $\psi_R{}^{(v/V)}$ from which, by (6),

$$p = 1 - \exp\left\{-\frac{\mathfrak{J}}{t} N\left[\frac{v}{V} N^{-1}\left(\frac{t}{\mathfrak{J}} \ln (1-P)^{-1}\right)\right]\right\}, \tag{26}$$

where $N^{-1}(x)$ denotes the inverse function.

The numeric solution to the problem at hand is, in effect, set forth in Fig. 1. Calculating the ordinate, the inverse, which is the corresponding single glimpse probability, is then found from the appropriate contour raised to the v/V power, and the new ordinate found. The remaining calculations are trivial.

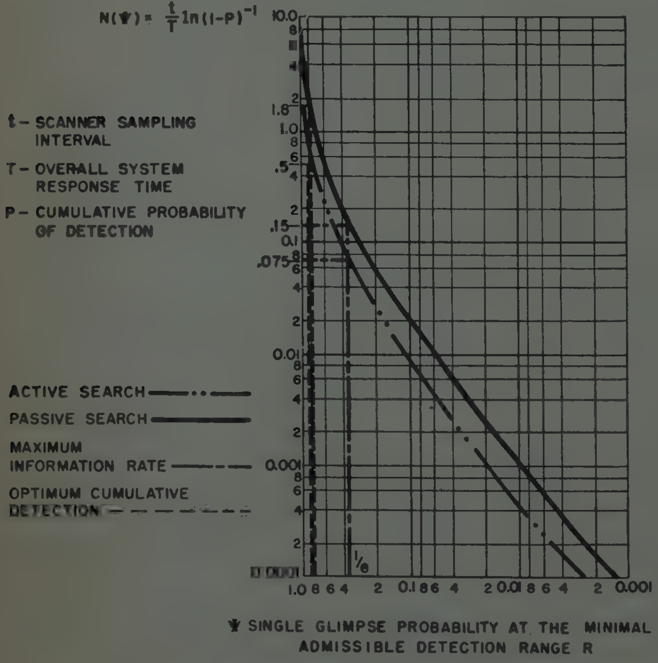

Fig. 1.

A representative numeric example will be illustrative of the manner of application and the pronounced dependence of the system upon V. To demonstrate the effect of a nominal increase in V of 0.1, assume $P=0.9$ for an active system characterized by a time constant to frame time ratio of 5. The inverse of $N(x)=\frac{1}{5}\ln(1-P)^{-1}$ is 0.71. Correcting for the velocity increase, the required single glimpse probability is 0.69^{-}, the reduction being caused by the fact that the desired detection range must increase with the closure rate. The corresponding value of N is 0.32 from which $p=1-\exp[-5(0.31)]=0.80$, which is a reasonably substantial degradation in view of the very moderate increase in relative velocity. A corresponding calculation for $P=0.5$ yields $p=0.3$, an even larger percentile decline in performance, whereas no change in the required cumulative detection probability is discernible if its initial value is 0.95. The preceding displays an important consideration in warning system design. In effect, detection performance must be good everywhere, or it is good nowhere, if the possible contact space is uniformly covered.

B. The Cumulative Influence of the Closure Rate

The conflict between the performance at various bearing angles and the requirements introduced by the bearing dependent frequency of potential collision contacts has been noted above. A useful measure of system effectiveness can be obtained by cascading the two phenomena.

In the analysis that follows, $\tilde{P}$ denotes the likelihood of such a contact during some differential time interval ϵ, and P denotes the probability of discerning the contact by the minimal range and responding by some appropriate action. Then, the likelihood of surviving beyond that interval is $P\,\tilde{P}+(1-\tilde{P})$. Noting that $\tilde{P}$ is of the form $\rho\epsilon V/T$, where ρ is a constant that reflects the particle density and T is the time duration of the flight, S, the likelihood of survival during some differential time interval with no collisions from some given bearing sector characterized by closure rates V, is given by

$$S = \lim_{\epsilon\to 0}\,[1-(1-P)\rho V\epsilon/T]^{T/\epsilon} = e^{-CV(1-P)}, \tag{27}$$

where C is a parameter that reflects environmental and system characteristics.

Plainly, S decays with increasing V since both factors of $V(1-P)$ are monotone increasing functions of V. It is reasonable to require equivalent performance in all feasible contact sectors, thereby leading to the idealized functional form $P=1-K^2/V$, if the exponential in (27) is to be constant. This is in obvious conflict with the functional behavior of what has been studied thus far and, indeed, is impossible for V sufficiently small, since P would then assume negative values. This latter contingency would not appear to be of concern in the case of astronautical vehicles in view of the anticipated particle velocities. In general, the factor $1-P$ is a far greater contributor to the variation with velocity of (27) than is the linear function V. Accordingly, the detection process will later be suboptimized by attempting to maintain $P(\beta)$ constant through variation of the search system dwell.

It will be well to accomplish some numerical analysis of the over-all degradation in terms of (27) before departing from this point. To demonstrate the consequences of a uniform rate of interrogation of the space volume, the earlier calculations will be extended. Let S denote the likelihood of adequate coverage over some sector characterized by collision closure rate V, subject to which the detection system yielded a cumulative probability of P. Then at another bearing angle whose corresponding parameters are v, p, the survival probability s is, from the above,

$$s = S^{v(1-p)/V(1-P)} = S^{v/V(1-p)V/v-1}.$$

It is apparent from the form of (28) that, as before, the degradations caused by an increased V are magnified as the initial likelihood of survival declines. On the other hand, the sensitivity to a modest increase in the ratio v/V declines with P, though it may be expected that S and P will simultaneously be high.

IV. SUBOPTIMIZATION IN THE CASE OF RANDOMLY MOVING PARTICLES

As indicated above, the distribution of dwell time will now be specified subject to the constraints that $P(V\mathfrak{Z})$ is to be maintained constant while the sweep time or search interval t is fixed. It is, of course, possible to investigate this problem without the latter constraint. However, sensor performance, computer bandwidth limitations in the data processing and decision-making apparatus would appear to establish the maximum rate at which information can be accepted for processing.

The desired result can be established without reference to the specific exponential form of the Rayleigh target. Since the signal integration time is inversely proportional to ω, the rate at which the angular sector is swept, the single glimpse probability will be assumed to be of the form $\psi(\rho\omega^{1/\nu})$ where ν has the same significance as before. One must then have

$$P = 1 - \exp\left[\frac{1}{V(\beta)t}\int_{V(\beta)\mathfrak{Z}}^{\infty} \ln\left\{1 - \psi[\rho\omega^{1/\nu}(\beta)]\right\} d\rho\right] \quad (29)$$

which, following the substitution $\rho\omega^{1/\nu}(\beta) = \phi$, may be written

$$P = 1 - \exp\left\{-\frac{\mathfrak{Z}}{t}\left[\frac{1}{\mathfrak{Z}V\omega^{1/\nu}}\int_{\mathfrak{Z}V\omega^{1/\nu}}^{\infty} \ln\,[1-\psi(\phi)]d\phi\right]\right\}. \quad (30)$$

It is apparent from the above that if $V\omega^{1/\nu}$ is maintained constant, P will be invariant with respect to the bearing angle. Alternately stated, the search dwell times should be so chosen that the integrated signal is proportional to the probability density of potential collision contacts raised to the $\nu/2$ power.

It is instructive to complete the analysis for the important case in which the search is conducted over the plane of motion of the observer. The vector diagram for collision immediately yields

$$0 = V^2 - 2vV\cos\beta - (v_p^2 - v^2), \quad (31)$$

where v, v_p denote, respectively, the absolute values of the velocity vectors of the observer and particle, and V denotes the relative velocity of approach. The preceding quadratic equation then yields

$$\begin{aligned} V &= v\cos\beta \pm \sqrt{(v_p^2 - v^2\sin^2\beta)} \\ &= v\cos\beta \pm \sqrt{\left[\left(v_p^2 - \frac{v^2}{2}\right) + \frac{v^2}{2}\cos 2\beta\right]}, \end{aligned} \quad (32)$$

which displays the well-known properties that for $v > v_p$ there are two solutions over the regions $0 \le \beta < \sin^{-1} v_p/v$, one at the upper boundary and none beyond. From the condition that the frame time t is specified and the preceding analysis which states that one must select $\omega^{-1} = K^{-1}V^{\nu}$, for $v \le v_p$, the case of greatest interest,

$$\begin{aligned} t &= \int_{-\pi}^{\pi} \frac{d\beta}{\omega} = \frac{2}{K}\int_0^{\pi} V^{\nu}(\beta)d\beta \\ &= \frac{2}{K}\int_0^{\pi} (v_p^2 + v^2\cos 2\beta)d\beta \\ &\quad \pm \frac{4v}{K}\int_0^{\pi} \cos\beta\sqrt{\left[\left(v_p^2 - \frac{v^2}{2}\right) + \frac{v^2}{2}\cos 2\beta\right]}d\beta \end{aligned} \quad (33)$$

for passive systems, and, for active search,

$$\begin{aligned} t &= \frac{2}{K}\int_0^{\pi} (v^4\cos 4\beta + 2v^2v_p^2\cos 2\beta + v_p^4)d\beta \\ &\quad \pm \frac{8v}{K}\int_0^{\pi} \cos\beta\{v^2\cos 2\beta + v_p^2\} \\ &\qquad \cdot\sqrt{\left[\left(v_p^2 - \frac{v^2}{2}\right) + \frac{v^2}{2}\cos 2\beta\right]}d\beta. \end{aligned} \quad (34)$$

Consideration of the symmetry and sense of the trigonometric functions show that integrals of $\cos\beta f(\cos 2\beta)$ over $0 \le \beta \le \pi$ are null. Accordingly, K is $2\pi v_p^{\nu}/t$ and, writing $r = v/v_p$,

$$\omega = \frac{2\pi}{t}\left(\frac{v_p}{V}\right)^{\nu} = \frac{2\pi}{t}\left[r\cos\beta + \sqrt{(1 - r^2\sin^2\beta)}\right]^{-\nu}, \quad (35)$$

where, as before, ν is 2 or 4 according to whether the search involves passive or active equipment. From (35) it is apparent that ω is a stronger function of β in the latter case. However, in both instances the functional dependence of the optimum scan rate upon the bearing angle declines with r, which is in accord with physical reasoning. The analysis in 3 dimensions proceeds in an identical manner, K being altered.

The above discussion established the required program for the distribution of the search effort subject to some specified sampling interval. To complete the analysis, it is necessary to develop a rational basis for the choice of the frequency with which the potential contact space is to be interrogated. It will now be shown that if, as has been done, only signal integration time is considered, a sampling interval exists which is optimum in the sense that it maximizes the cumulative probability of detection at some stipulated desired range. Since, as has been noted, the latter is dependent only upon the anticipated closure rate and the system reaction time, and since, all else fixed, the signal level $S(t)$ is a monotone increasing function of the integration time, the cumulative probability of detection assumes the form

$$\begin{aligned} P(t) &= 1 - \exp\left[\frac{1}{Vt}\int_{V\mathfrak{J}}^{\infty} \ln\left(1 - \phi(t)e^{-a\rho^{\nu}/S}\right)d\rho\right] \\ &= 1 - \exp\left[\frac{1}{V}\,\frac{F(t)}{t}\right]. \end{aligned} \tag{36}$$

$\phi(t)$ is a generalized degradation factor that reflects the efficiency of the data handling subsystem. Explicitly, it measures the probability that cognizance is taken of the individual sensor outputs and is therefore a positive function, bounded above by unity.

Only the behavior of $F(t)/t$ need be investigated. The latter ratio may be exhibited in the form,

$$\begin{aligned} \frac{F(t)}{t} &= -\frac{1}{t}\int_{V\mathfrak{J}}^{\infty}\left\{\sum_{\lambda=1}^{\infty}\frac{\phi^{\lambda}(t)}{\lambda}\,e^{-\lambda a\rho^{\nu}/S(t)}\right\}d\rho \\ &= -\frac{1}{t}\sum_{\lambda=1}^{\infty}\frac{\phi^{\lambda}}{\lambda}\int_{V\mathfrak{J}}^{\infty}e^{-\lambda a\rho^{\nu}/S}d\rho \\ &= -\frac{1}{\nu a^{1/\nu}}\sum_{\lambda=1}^{\infty}\frac{1}{\lambda^{1+1/\nu}} \\ &\quad\cdot\left\{\phi^{\lambda}\,\frac{S^{1/\nu}}{t}\int_{\lambda a/S(V\mathfrak{J})^{\nu}}^{\infty}e^{-x}x^{1/\nu-1}dx\right\}. \end{aligned} \tag{37}$$

For t sufficiently small, the bracketed factor $g(t)$ in the generating term of the above infinite series obeys the train of inequalities,

$$\begin{aligned} 0 \le g(t) &\le \frac{S}{t}\left[\lambda a(V\mathfrak{J})^{\nu}\right]^{1/\nu-1}\int_{\lambda a/S(V\mathfrak{J})^{\nu}}^{\infty}e^{-x}dx \\ &= \frac{S}{t}\left[\lambda a(V\mathfrak{J})^{\nu}\right]^{1/\nu-1}e^{-(\lambda a/S)(V\mathfrak{J})^{\nu}}. \end{aligned} \tag{38}$$

Ideally, the mean signal to root-mean-square noise voltage ratio is proportional to $\sqrt{t}$, in which case $S(t)$ may be taken as proportional to t. The upper bound then vanishes with t because of the behavior of the exponential factor. Indeed, there appears to be no physically meaningful sensor and signal integration process that may be characterized by an $S(t)$ that has a zero at the origin of sufficiently low order to yield other than a zero limit for $F(t)/t$ and, in turn, $P(t)$, which displays the same asymptotic behavior for t small.

$$\begin{aligned} \left|\frac{F(t)}{t}\right| &= \frac{1}{t\nu a^{1/\nu}}\sum_{\lambda=1}^{\infty}\frac{\phi^{\lambda}S^{1/\nu}}{\lambda^{1+1/\nu}}\left(\int_0^{\infty} - \int_0^{\lambda a/S(V\mathfrak{J})^{\nu}}\right)e^{-x}x^{1/\nu-1}dx \\ &= \frac{1}{t\nu a^{1/\nu}}\sum_{\lambda=1}^{\infty}\frac{\phi^{\lambda}}{\lambda^{1+1/\nu}}\left\{\Gamma\left(\frac{1}{\nu}\right) - \gamma\left(\frac{1}{\nu}, \frac{\lambda a}{S}(V\mathfrak{J})^{\nu}\right)\right\} \\ &< \frac{\Gamma\left(\frac{1}{\nu}\right)S^{1/\nu}}{t\nu a^{1/\nu}}\,\zeta\left(1+\frac{1}{\nu}\right), \end{aligned} \tag{39}$$

where $\Gamma(n)$ and $\gamma(n, x)$ represent the complete and incomplete gamma functions, respectively, in the notation of Legendre, and $\zeta(x)$ is the Riemann zeta function. In general, t is far more rapidly increasing than is $S^{1/\nu}(t)$ which, for many classes of integrators, is bounded above. Accordingly, $S^{1/\nu}(t)/t$ vanishes in the limit as the sampling interval becomes indefinitely large, as in the case of the idealized integrator previously cited. In that instance the result follows immediately since the variable appears only in the denominator which contains the factor $t^{1-1/\nu}$, the exponent being positive. Since $P(t)$ is positive, nonconstant, and vanishes in the limit at both the origin and the point at infinity, it must have at least one maximum, which was to be shown.

The uniqueness of the optimum scan interval could not be established with the highly generalized detection processes considered above. However, if an idealized signal integration process and an essentially unsaturated data handling subsystem are assumed, in this, the most important case, the uniqueness can be readily established. Differentiation of $F(t)/t$ with respect to the sampling interval yields

$$\begin{aligned} &\frac{\partial}{\partial t}\left(\frac{F(t)}{t}\right) \\ &\quad = -\frac{F(t)}{t^2} + \frac{1}{t}\int_{V\mathfrak{J}}^{\infty}\frac{\partial}{\partial t}\ln\left(1 - e^{-a\rho^{\nu}/t}\right)d\rho. \end{aligned} \tag{40}$$

The partial derivative of the integrand is

$$\begin{aligned} &-\frac{a}{t^2}\,\rho^{\nu}e^{-a\rho^{\nu}/t}/\left(1 - e^{-a\rho^{\nu}/t}\right) \\ &\qquad = -\frac{\rho}{\nu t}\,\frac{\partial}{\partial\rho}\ln\left(1 - e^{-a\rho^{\nu}/t}\right), \end{aligned} \tag{41}$$

from which integration by parts yields

$$\begin{aligned} &\frac{\partial}{\partial t}\left(\frac{F(t)}{t}\right) \\ &\quad = -\frac{1}{\nu t^2}\left\{(\nu - 1)F(t) - V\mathfrak{J}\ln\left(1 - e^{-a(V\mathfrak{J})^{\nu}/t}\right)\right\}. \end{aligned} \tag{42}$$

From the earlier discussion, the possible zeros at the origin and the point at infinity are not of interest, so that the desired extremals are the finite zeros of the numerator at other than the origin.

Accordingly,

$$\frac{F(t)}{Vt} = \frac{\mathfrak{J}}{(\nu - 1)t} \ln \left(1 - e^{-a(V\mathfrak{J})^{\nu}/t}\right) \tag{43}$$

and

$$P = 1 - \left(1 - e^{-a(V\mathfrak{J})^{\nu}/t}\right)^{\mathfrak{J}/(\nu-1)t} \tag{44}$$

in the vicinity of a maximum (or a minimum). Verbalized, if the scan rate of a detector has been properly chosen, the cumulative detection probability at the range at which performance is optimized is the same as if $1/(\nu-1)$ of the remaining possible scans were taken at the single-glimpse probability which obtains at the range in question. Writing (44) in the form,

$$\frac{t}{\mathfrak{J}} \ln (1 - P)^{-1} = \frac{1}{\nu - 1} \ln (1 - \psi)^{-1}, \tag{45}$$

the desired extremals may be readily found through graphic solution by constructing the right-hand function on Fig. 1. The resultant optima for both passive and active systems, which are very nearly identical, as well as the value of $N(\psi)$ for the maximum information rate choice of the sampling interval, are displayed.

Unless optimization in the above sense is being accomplished in the vicinity of reasonably high signal levels, the required signal integration may lead to unacceptably low information rates over the scanned space since the required single glimpse probabilities are of the order of 0.8. This is particularly true in the case of the low signal amplitude, hypervelocity closure rate environment under consideration, leading thereby to the conclusion that it may not be possible to so optimize astronautical vehicle search systems. However, the information content derived during an individual interrogation improves with increased dwell times, implying that the optimal search rates in this application are intermediate between the extremes dictated by tracking considerations and detection in the classical sense.

V. PARTICULARIZATION OF SEARCH DURING MOTION UNDER A CENTRAL FORCE FIELD

Let $\mathbf{P}$, $\mathbf{p}$ represent, respectively, the position vectors of the astronautical vehicle and another particle, both moving freely under the influence of a central force field, the source of which is located at the origin of vectors. It will be further assumed that M, the mass of the latter body, is extremely large with respect to that of either of the others and that there is no discernible gravitational interaction between the astronaut and the other particle. As is apparent, the preceding conditions characterize the environmental kinematics of space.

From the above constraints, by use of the dot superscript notation to indicate time derivatives, one may write for the pertinent differential equations of motion,

$$\ddot{\mathbf{P}} = \mathbf{P}f(|\,\mathbf{P}\,|), \qquad \ddot{\mathbf{p}} = \mathbf{p}f(|\,\mathbf{p}\,|), \tag{46}$$

where $f(|\,\mathbf{A}|)$ may be any well-behaved function of the absolute value of the position vector. It may be readily shown that despite the admissible approximation of uniform rectilinear motion in the vicinity of contact a collision course, defined in the conventional manner in which the relative velocity vector is coincident with the relative positional vector, is impossible unless both objects are on the degenerate linear orbit containing the center of force. Vectorially, this corresponds to stating that their cross product must vanish, or

$$0 = (\dot{\mathbf{p}} - \dot{\mathbf{P}}) \times (\mathbf{p} - \mathbf{P}), \tag{47}$$

which may be rearranged in the form,

$$\dot{\mathbf{p}} \times \mathbf{P} + \dot{\mathbf{P}} \times \mathbf{p} = \dot{\mathbf{p}} \times \mathbf{p} + \dot{\mathbf{P}} \times \mathbf{P}. \tag{48}$$

However, the derivative of the right side of the above must vanish identically since, as is well known, the areas swept per unit period of time by the radii vectors are constant. Differentiation of the left side of (48) leads to

$$\{\ddot{\mathbf{p}} \times \mathbf{P} + \ddot{\mathbf{P}} \times \mathbf{p}\} + \{\dot{\mathbf{p}} \times \dot{\mathbf{P}} + \dot{\mathbf{P}} \times \dot{\mathbf{p}}\} = 0, \tag{49}$$

where, plainly, the second bracketed quantity vanishes. Applying (46) to the remaining quantity, one must have

$$[f(|\,\mathbf{p}\,|) - f(|\,\mathbf{P}\,|)]\mathbf{p} \times \mathbf{P} = 0. \tag{50}$$

The scalar factor has, at best, isolated zeros unless the orbits are identical and, if not circular, they must be concurrently swept. Accordingly, the vector product $\mathbf{p}\times\mathbf{P}$ must be null for a finite interval preceding the contact, or $\mathbf{p}=m\mathbf{P}$, where m is some scalar quantity. Replacement of $\mathbf{p}$ and $\dot{\mathbf{p}}$ in (47) by the appropriate relationships in $\mathbf{P}$ and $\dot{\mathbf{P}}$, or the reverse, leads to

$$\begin{aligned} 0 &= [(\dot{m}\mathbf{P} + m\dot{\mathbf{P}}) - \dot{\mathbf{P}}] \times [(m\mathbf{P}) - \mathbf{P}] \\ &= (m - 1)^2\dot{\mathbf{P}} \times \mathbf{P} \end{aligned}$$

and

$$\begin{aligned} 0 &= \left[\dot{\mathbf{p}} - \frac{1}{m}\left(\frac{\dot{m}}{m}\mathbf{p} - \dot{\mathbf{p}}\right)\right] \times \left[\mathbf{p} - \left(\frac{1}{m}\mathbf{p}\right)\right] \\ &= \left(1 - \frac{1}{m^2}\right)\dot{\mathbf{p}} \times \mathbf{p}. \end{aligned} \tag{51}$$

Exclusive of the trivial case in which m is unity and the vectors are identical, the velocity and position vectors of each particle must be parallel, which was to be shown. Certainly, such trajectories are improbable since fuel consumption considerations dictate minimal energy paths such as the Hohmann cotangential ellipses.

However, despite the fact that in a precise sense a collision course is impossible except in a trivial case, certain general conclusions may be drawn regarding a feasible contact zone. It will now be shown that particles approaching with an essentially invariant bearing angle β must have relative trajectories that lie on a surface that is characterized by reasonable navigational system inputs relating to the motion of the astronautical vehicle in the central force field. Further, the surface will be seen to be independent of the specific nature of the law governing the intensity of the force field, though this observation is largely of academic interest since the inverse square law appears to fail only over regions that are either too close or too distant to the center of force to be of consequence. What is important, however, is that, referencing the earlier theory, confinement of the search to this limited region rather than interrogating omnidirectionally makes major improvements in detector performance possible.

Recognizing the essentially rectilinear nature of the motions of the astronautical vehicle and the contact, the left side of

$$-\dot{\tilde{\beta}}\sin\tilde{\beta} = \frac{d}{dt}\left\{\frac{[(\dot{\mathbf{p}}-\dot{\mathbf{P}})\cdot\dot{\mathbf{P}}]}{|\dot{\mathbf{p}}-\dot{\mathbf{P}}|\,|\dot{\mathbf{P}}|}\right\} \tag{52}$$

must vanish as the condition of warning. From the definition of the scalar product, $\tilde{\beta}$ is the supplement of the bearing angle. Accomplishing the indicated differentiation, (52) becomes

$$\begin{aligned}-\dot{\tilde{\beta}}\sin\tilde{\beta} = {} & \frac{[(\ddot{\mathbf{p}}-\ddot{\mathbf{P}})\cdot\dot{\mathbf{P}}]+[(\dot{\mathbf{p}}-\dot{\mathbf{P}})\cdot\ddot{\mathbf{P}}]}{|\dot{\mathbf{p}}-\dot{\mathbf{P}}|\,|\dot{\mathbf{P}}|} \\ & - \frac{[(\dot{\mathbf{p}}-\dot{\mathbf{P}})\cdot\dot{\mathbf{P}}]}{|\dot{\mathbf{p}}-\dot{\mathbf{P}}|^2|\dot{\mathbf{P}}|^2} \\ & \cdot\left\{|\dot{\mathbf{p}}-\dot{\mathbf{P}}|\,\frac{d}{dt}\,|\dot{\mathbf{P}}| + |\dot{\mathbf{P}}|\,\frac{d}{dt}\,|\dot{\mathbf{p}}-\dot{\mathbf{P}}|\right\}.\end{aligned} \tag{53}$$

Utilizing the fact that the derivative of the absolute value of any non-null vector $\mathbf{A}$ is $(\dot{\mathbf{A}}\cdot\mathbf{A})/|\mathbf{A}|$, (53) may be replaced by

$$\begin{aligned}-\dot{\tilde{\beta}}\sin\tilde{\beta} = {} & \frac{[(\ddot{\mathbf{p}}-\ddot{\mathbf{P}})\cdot\dot{\mathbf{P}}]+[(\dot{\mathbf{p}}-\dot{\mathbf{P}})\cdot\ddot{\mathbf{P}}]}{|\dot{\mathbf{p}}-\dot{\mathbf{P}}|\,|\dot{\mathbf{P}}|} \\ & - \frac{[(\dot{\mathbf{p}}-\dot{\mathbf{P}})\cdot\dot{\mathbf{P}}]}{|\dot{\mathbf{p}}-\dot{\mathbf{P}}|^2|\dot{\mathbf{p}}|^2}\left\{\frac{|\dot{\mathbf{p}}-\dot{\mathbf{P}}|}{|\dot{\mathbf{P}}|}(\dot{\mathbf{P}}\cdot\ddot{\mathbf{P}})\right. \\ & \left. + \frac{|\ddot{\mathbf{P}}|}{|\dot{\mathbf{p}}-\dot{\mathbf{P}}|}[(\dot{\mathbf{p}}-\dot{\mathbf{P}})\cdot(\ddot{\mathbf{p}}-\ddot{\mathbf{P}})]\right\}\end{aligned} \tag{54}$$

and, finally, from (46), for all objects moving independently in a communal central force field,

$$\begin{aligned}-\dot{\tilde{\beta}}\sin\tilde{\beta} = {} & \{[(f(|\mathbf{p}|)\mathbf{p}-f(|\mathbf{P}|)\mathbf{P})\cdot\mathbf{P}] \\ & + f(|\mathbf{P}|)[(\dot{\mathbf{p}}-\dot{\mathbf{P}})\cdot\mathbf{P}]\}\,|\dot{\mathbf{p}}-\dot{\mathbf{P}}|^{-1}|\dot{\mathbf{P}}|^{-1} \\ & - \frac{[(\dot{\mathbf{p}}-\dot{\mathbf{P}})\cdot\dot{\mathbf{P}}]}{|\dot{\mathbf{p}}-\dot{\mathbf{P}}|^2|\dot{\mathbf{P}}|^2}\left\{\frac{f(|\mathbf{P}|)\,|\dot{\mathbf{p}}-\dot{\mathbf{P}}|}{|\dot{\mathbf{P}}|}(\dot{\mathbf{P}}\cdot\mathbf{P})\right. \\ & + \frac{|\dot{\mathbf{P}}|}{|\dot{\mathbf{p}}-\dot{\mathbf{P}}|}[(\dot{\mathbf{p}}-\dot{\mathbf{P}})\cdot(f|\mathbf{p}|)\mathbf{p} \\ & \left. - f(|\mathbf{P}|)\mathbf{P}]\right\}.\end{aligned} \tag{55}$$

In the vicinity of impact, $\dot{\tilde{\beta}}\rightarrow 0$ and $\mathbf{p}\rightarrow\mathbf{P}$, so that

$$\begin{aligned}0 = {} & \frac{f(|\mathbf{P}|)}{|\dot{\mathbf{p}}-\dot{\mathbf{P}}|\,|\dot{\mathbf{P}}|}\left\{[(\dot{\mathbf{p}}-\dot{\mathbf{P}})\cdot\mathbf{P}]\right. \\ & \left. - \frac{[(\dot{\mathbf{p}}-\dot{\mathbf{P}})\cdot\dot{\mathbf{P}}](\dot{\mathbf{P}}\cdot\mathbf{P})}{|\dot{\mathbf{P}}|^2}\right\}.\end{aligned} \tag{56}$$

However, this is only possible if the bracketed expression vanishes. The latter condition may be exhibited in the form,

$$\frac{[(\dot{\mathbf{p}}-\dot{\mathbf{P}})\cdot\dot{\mathbf{P}}]}{|\dot{\mathbf{p}}-\dot{\mathbf{P}}|\,|\mathbf{P}|} = \frac{[(\dot{\mathbf{p}}-\dot{\mathbf{P}})\cdot\dot{\mathbf{P}}]}{|\dot{\mathbf{p}}-\dot{\mathbf{P}}|\,|\dot{\mathbf{P}}|}\;\frac{(\dot{\mathbf{P}}\cdot\mathbf{P})}{|\dot{\mathbf{P}}|\,|\mathbf{P}|}$$

or

$$\cos\mu = \cos\tilde{\beta}\cos\gamma, \tag{57}$$

where μ is the angle between the relative velocity and astronaut positional vectors. γ, the angle between the velocity and positional vectors of the astronaut, may be presumed known. The relationship (57) defines a surface over which, viewed by the astronaut, the potential collision contact must make its relative terminal approach.

VI. CONCLUSIONS

1) Programmed space distribution of system search effort as a function of the geometry of the distribution of potential contacts and the anticipated relative closure rate is mandatory.

2) The probability of warning by some specified range of an impending collision markedly declines with increasing closure rates while, fixing over-all system response time, detection range requirements increase. Further, random particle distributions bias the frequency of potential collision contacts in the direction of the higher closure rates so that survival probability is seriously degraded if the observer space is omnidirectionally and uniformly interrogated by the search system.

3) All else equal, the number of successful interrogations per unit time interval of a Rayleigh contact by a continuously scanning system is maximized by choosing signal integration times that yield a single glimpse probability of $1/e$. The probability of receiving at least one

return in a given time interval is maximized by choosing integration times that yield a single glimpse probability of 1/2. Scan rates are thereby implied by automatic data handling needs that are substantially higher than those dictated by conventional criteria of optimal detection performance which conflict with the short system reaction time requirements.

4) If the rate at which observer space is interrogated has been chosen so as to yield the greatest attainable cumulative probability of detection at some specified range, the latter probability is that which would result if $1/(\nu-1)$ of the remaining scans preceding collision were made at the single glimpse probability which obtains at the range in question, ν being 4 or 2 according to whether the search system is active or passive.

5) If detection performance requirements are defined by over-all system response time so that the minimal admissible warning range is the product of the latter and the anticipated closure rate, and the closure rate is definable over the observer space, the probability of acquiring the contact by that critical range may be maintained constant over the potential contact space by choosing search system dwell times over each sector that are proportional to the $1/\nu$th power of the characteristic closure rate.

6) If the mutual gravitational interaction between a particle and an astronautical vehicle moving in a communal central force field is negligible, as viewed by the astronaut the relative approach path of the former is essentially confined to a surface defined by reasonable navigational system inputs.

7) Though generally a complex function of the signal amplitude, if the latter is very low the weighted mean acquisition range exhibits the idealized $1/\nu$th power dependence upon signal power, where ν has the same meaning as above.

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An Inertial Guidance System Field Test Program for a Ballistic Missile Weapon System*

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Summary—**A field test program for an inertial guidance system can easily extend from coast to coast and will require large numbers of experienced personnel and equipments. Included in the field test program are *Sled Tests*, *Captive Tests*, *Flight Test*, *Operational Tests* and *Contractor Liaison*. These areas of tests are described and objectives are reviewed and discussed.**

IT is recognized that the actual use of a weapon system in its operational environment is the only conclusive test of that weapon system. However, an intelligently planned, integrated and executed test program can establish a high degree of confidence in the operational performance of a given system design. Therefore, a test program must be so planned and executed as to fully qualify a given system design.

The concept of a design verification program places extreme emphasis upon the following test concepts.

1) Dynamic operating tests on a system is the goal in all phases of the test program.
2) Primary emphasis is placed upon the testing of the most inclusive assembly available. For example, module testing is emphasized until components become available for test; then component testing is emphasized until systems become available for test, at which time the primary emphasis is shifted to system testing. Module and component testing are then carried on merely to complement system testing by further isolating malfunctions revealed in the more inclusive system tests.
3) Design verification is thoroughly integrated and is planned so that all phases of the test program complement and support each other.

The test program supplements the reliability program in determining component and system reliability using mean time to failure as a yardstick.

A complete test program outline includes the following:

I. In Plant Tests
  A. Engineering Tests
    1) Gyro and Accelerometer Study Program
    2) REAC Computer Analysis Program
    3) Breadboard Tests
    4) Product Improvement
  B. Environmental Tests
    1) Verification Tests
    2) Flight Proofing Tests
    3) Evaluation Tests
    4) Qualification Tests
II. Field Tests
  A. Sled Tests
  B. Contractor Liaison
  C. Captive Tests
  D. Flight Tests
  E. Operational Tests

The general objectives that must be accomplished during each phase of the field tests follow.

SLED TESTS

During the entire R and D effort, a rocket sled test program will complement the environmental and flight proofing programs.

The sled environment provides jerk inputs which are difficult to obtain by other means. Sled testing also allows simultaneous operation of all axes of the inertial measurement unit during application of the combined acceleration-vibration environment of the sled. This allows evaluation of interaction or cross-coupling among the inertial measurement units axes.

Prior to the first flight test of a complete guidance system, the sled test will have demonstrated system operation in a combined acceleration and vibration environment, complementing the in-plant flight-proofing program.

Information obtained from the sled tests during this phase is much more comprehensive than the information obtained from the early missile flights. This greater capability is accomplished by provision for telemetering a greater number of signals and almost unlimited flexibility in choice of information to be monitored. Recoverability of the equipment used on sled tests and subsequent evaluation provide valuable information not easily obtainable by other methods.

The main phases of a sled test program are now listed in chronological order.

Evaluation of Sled Environment

This phase obtains sled environment information while checking out the basic sled equipment and familiarizing personnel with test procedures. The suitability of sled-borne power supplies and of equipment mounting brackets are demonstrated.

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Evaluation of Signal Conditioner Component and Telemetry

The signal conditioner and telemetry stability in the severe sled environment are demonstrated. The reliability and mechanical integrity of the conditioner and telemetry are demonstrated.

Accelerometer Assembly Tests

The effect of the sled environment on various accelerometer constants is evaluated. In addition, the associated instrumentation, recording and data reduction techniques are evaluated during these tests.

Qualitative Erection and Stabilization Tests

This phase demonstrates functional operation of the equipment and the applicable portion of the countdown procedures are evaluated.

Velocity Measurement Tests

Correlation of system and velocity measurement is the objective of this phase. Velocity measurement errors and stabilization errors are isolated.

Qualitative System Tests

The complete system is functionally demonstrated during this phase. The complete countdown procedure is evaluated.

Quantitative System Evaluation

This phase consists of running test problems which are used to accomplish cutoff or other signals demonstrating computer operation.

Flight Support Tests

Areas indicated for critical evaluation by flight test results are investigated. These tests can take priority over others at any time if the flight test program is in progress.

Marginal Conditions Tests

The effects of non-nominal calibrations and/or power supplies are investigated. Included in these tests are the effects of power supply transients on the system operation.

CONTRACTOR LIAISON

In order to accomplish the over-all objectives of a successful weapon system, a great need exists for a coordinated effort during the entire test program by all the contractors concerned. In line with this objective, the associate contractor must plan to have a sound engineering group located at the air frame contractor's facility.

The airframe contractor activities which require support will include the following areas:

a) Missile engineering tests (integration tests)
b) Acceptance tests
  1) Receiving and inspection
  2) Installation and acceptance testing
c) Captive tests
d) Laboratory
  The lab will support acceptance tests, missile engineering tests and captive tests.

Integration Tests

Missile engineering tests will consist of laboratory tests of the integration of the guidance equipment into the missile system. The basic compatibility tests must be completed before commencement of the flight test program.

The primary objective of the missile engineering test program is to evaluate compatibility, functional integration tests, and engineering evaluation tests of the guidance set, missile, and missile subsystems.

Specific test objectives are:

1) Verify that the system physically can be installed into the missile.
2) Demonstrate the compatibility of the ground support equipment with the missile.
3) Demonstrate the compatibility of the inertial guidance set with other subsystems.
4) Evaluate the system operation with missile power supplies.
5) Evaluate the effects of power transfer and power transients on system operation.
6) Evaluate system performance with the missile cabling and power to verify that ground loops do not exist and noise levels and interference from other subsystems do not affect guidance system operation.

Acceptance Testing

Following a complete inertial guidance set calibration, functional tests, and acceptance at the contractor plant, inertial guidance sets will be transported to the airframe manufacturer for inspection, checkout, installation, and missile acceptance checks.

Captive Tests

Static firings of complete missiles are accomplished by the airframe contractor at a captive test facility. Captive tests can be accomplished in two ways. They can be accomplished with the static firing of production missiles after complete functional integration tests have been completed, or a test program can be implemented that will static fire one or several missiles many times.

The tests are designed primarily to increase confidence in launches and to improve reliability. The later tests accomplish this over a shorter period of time. The former program is better as the configuration of the missile changes, and the captive test is an immediate positive test of that configuration.

FLIGHT TEST PROGRAM

The ultimate design verification and weapon system evaluation can be obtained only from the results of flight tests. In view of this, a strong and early program must be planned.

The important phase of the flight test program will sequentially accomplish the following objectives:

1) Permit any design modification dictated by flight test data to be accomplished prior to an extensive production effort toward operational equipment.
2) Provide actual environment data, making possible a review of the environments assumed for purpose of design.
3) Demonstrate the mutual compatibility of the operation and function of the inertial guidance equipment under actual operational conditions.
4) Make possible the evaluation of the operation and function of the equipment under actual operation conditions.
5) Demonstrate the capability of the weapon system, and its subsystems, to perform the mission for which it is intended.

In general, flight tests evaluate the ability of the inertial guidance equipment to guide the missile to the cutoff point to effect the required ballistic trajectory. Passenger flights of components can in many cases provide a quantitative and qualitative evaluation of the components prior to actual inertially guided flight tests. Passive flights are not required if early passenger flights, laboratory, sled, flight proofing and other environmental testing are accomplished prior to system flight testing.

Airborne instrumentation must be complete enough so that performance and accuracy data will be obtained on any flight. Accuracy should be an objective of every flight commencing with the first flight. Power supply data should be available on all flights throughout the program to insure proper interpretation of flight data.

The primary intent of the flight tests is to determine the accuracy and the operating characteristics of the inertial guidance set during missile flight. The operating environment of the inertial guidance equipment is determined on several of the early flights.

Data obtained from each flight is analyzed, and the results are transmitted to the design engineers for reevaluation of the individual assemblies.

Specific flight test objectives.

1) Compatibility of the ground support equipment, missile, launch area and inertial guidance system.
2) Ability of the stabilization servo loops to isolate the inertial package from the base motion under dynamic conditions.
3) Accuracy of the accelerometer loops under dynamic conditions.
4) Accuracy of the computing elements in generating the correct steering, engine cutoff and prearm signals from the measured accelerations and the stored or programmed data inputs.
5) The accuracy, stability, reliability, performance, etc., of the power supplies before and during powered flight.
6) Accuracy of the inertial guidance system, based on a comparison of velocity data and external data determination of missile position throughout the flight profile and at the time of cutoff.
7) Determination of the operating environment.
8) General functional operation of erection and ground support equipment under field conditions.
9) Verification of computer operation, computer coefficients and other launcher-target parameters.
10) Evaluation of the staging transients and their effect on the operation of the inertial guidance set.

As the test program progresses, situations will arise which will show a need for additions or alterations to flight objectives. The program must be flexible so that the changes in objectives will be a simple matter. The associated equipment, such as the signal conditioner, must allow for immediate changes in telemetry channel assignment, if required in accordance with changes of objectives.

OPERATIONAL TESTS

It is recognized that the actual use of a weapon system in its operational environment is the only conclusive test of a weapon system. It is the intent of the operational tests to approach this environment as closely as possible.

The purpose of the operational tests is to demonstrate the existence of an operational weapon system. As many as possible of the events involved in the complete factory-to-launch sequence are tested. The primary objectives of the program are as follows:

1) Compatibility checks between the missile, ground support equipment and the operational missile launcher.
2) Evaluation of the ability of the handling and/or transportation and installation equipment to perform its task properly with adequate protection, ease, safety for the operating personnel, and with maximum efficiency in an operational environment.
3) Evaluation of the periodic checkout equipment, and procedures to ascertain its ability to verify that the weapon system is in the required state of readiness.
4) Evaluation of the operational countdown procedure and investigation of the possibility of decreasing the readiness time.
5) Evaluation of the procedure and time required for removal and installation of an inertial guidance set

and to return the missile to readiness condition.
6) Evaluation of launches which are accomplished by operational personnel and procedures.

Conclusion

It can be seen from the test program presented above that a large part of a complete test program must be accomplished in the field. This results in an extensive field test program that can extend from one end of the country to the other and also requires considerable numbers of experienced personnel and equipment. There are areas included in the field testing program which apparently could be adequately evaluated in the contractor's plant. Admittedly, there are portions of these areas that can be evaluated in the plant, and it is quite possible that they would receive a better evaluation there because of the high grade of personnel and facilities available. However, to make a completely thorough evaluation, field testing is still required. There are several reasons why extensive field testing is absolutely essential. Probably the most significant reason is that the exact operational conditions cannot be duplicated in the contractor's plant and the only conclusive test is the actual use of the weapon system in its operational environment.

A Pragmatic Approach to Space Communication*

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Summary—**Examination of the fundamental limitations on modulation efficiency and present state-of-the-art practices in transmitter design, solar cell power sources, receiver design and antenna design shows that communication with large information rates is possible throughout the solar system but that practical interstellar communication is not now feasible. Two representative telemetry systems are discussed, the FM/PM analog Microlock system and the digital Telebit system.**

Introduction

THE paper which follows attempts to establish the realistic limits which the state of the art of present day components places upon the transmission of information from space vehicles.

The fundamental problem of space communication is effective energy; that is, the energy necessary to transmit the required information over the distances characteristic of the various space missions. More specifically, it is the ratio of the thermal energy in the bandwidth required to transmit the desired information per unit time to the effective transmitted energy. Thus, the fundamental parameter of space communications is the energy per bit of information, and the fundamental limitation is ultimately that of the effective energy or the effective power of the system.

This basic limitation can be divided into two classes of problems. The first is the generation, transmission, and reception of energy. The second is determining the most effective method of modulating this energy to transmit the required information.

The information is determined by the particular space mission and the requirements of the experiments or experimenters. The rate of transmission of information is determined by the amount of data processing which is carried out in the satellite itself. Thus, the power in low-altitude earth satellites is sufficiently high and the distance is small enough so that sophistication in either modulation techniques or in data processing within the satellite is not essential. On the other hand, the extremely narrow bandwidth characteristic of interstellar probes requires a very highly sophisticated data processing in the satellite itself if appreciable information is to be transmitted from the experiments. From another aspect, the rapid motion of low-altitude earth satellites requires ground antennas of low directivity and a very large number of ground stations if continuous coverage is to be achieved. At the other extreme, high-altitude satellites for lunar probes or interplanetary probes require no more than three ground antennas for complete coverage, and since the motion of the satellite relative to the earth's surface is slow, these antennas can be very large.

Despite the wide variety of problems associated with each of the space missions, a broader view shows that available components impose a common limitation affecting each type of service. Thus, the power supply, the power amplifier, temperature control, the antennas, both in the air and on the ground, and the initial stages of the receiver on the ground are critical energy-determining components. In addition, there is the limitation on the efficiency of the modulation system imposed by available coding and decoding techniques. However, since the characteristics of these components and techniques are known, it is possible to trade off the component characteristic, the method of modulation, and the requirements of data processing to design an optimum system for any particular mission.

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The following sections will outline the parameters involved in energy transmission, the problems associated with the encoding and transmission of information, and will apply these results to four representative space missions.

ENERGY

As far as space communication is concerned, the problem of providing sufficient energy may be divided between the amount of energy available in the space vehicle, the degree to which that energy can be directed toward the ground receiver, and the efficiency of the receiving system in detecting this energy in the presence of noise.

Satellite Transmitting System

In the space vehicle, both the transmitted power and the total available energy are the important communication parameters. The effective power available from the space vehicle may be increased, 1) by increasing the transmitter power, 2) by increasing the transmitter efficiency, or 3) by increasing the antenna gain.

To increase transmitter power, it is necessary to provide a larger energy source, larger components, and heavier and more complex temperature control mechanisms. All of this results in an almost linear increase in payload weight with increasing power. The energy supply may be chemical, thermal, or nuclear in origin. The most efficient sources of chemical energy in terms of watt-hours per pound appear to be, at the present time, the various forms of the silver cadmium cells. Since this energy lies between 35 and 50 watt-hours per pound, the use of these cells as the primary source of energy for space vehicles is severely limited.

For earth and interplanetary satellites, it is possible to make use of the thermal energy of the sun, either by means of solar cells or thermocouples to provide a source of energy of relatively low-power capability but of very large total magnitude. At the present state of the art, it is apparent that for vehicles having lifetimes greater than two days, the solar cell is superior to a chemical battery in terms of weight for a given total energy.

Nuclear sources of electrical energy are in a very early stage of development, and it is not possible at this time to define precisely the energy per pound that will be available when their development is complete. Early estimates indicate some tens of kilowatt-hours per pound as being available. For missions far from the sun, such as flights to Pluto or to interstellar space, it appears that nuclear fission sources of energy are the only practical sources likely to be available within the next ten years.

Another problem to be considered is the efficiency with which this energy is converted into radio waves. At the present time it appears possible to achieve efficiencies between 50 and 80 per cent for transmitters below 1000 megacycles, and efficiencies on the order of 10 to 50 per cent for transmitters above 1000 megacycles. Moreover, since radio transmitters have been studied for a long period of time, major improvements in the operating efficiency are unlikely. The unused energy is heat and must be dissipated if the transmitter is not to be destroyed. Although the heat that can be radiated from the surface of the space vehicle depends upon its orientation with respect to the sun, for areas which are not exposed to direct radiation from planets or from the sun, it is possible to dissipate passively more than 300 watts per square foot.

Two factors then limit the amount of power that can be generated and transmitted from the space vehicle. First, the energy available, which is dependent upon the weight that can be allocated to the energy source, and the heat energy which can be dissipated. This depends in part upon the area available, that is, the size of the space vehicle, and upon the weight which can be allocated.

The effective power transmitted depends, in addition, upon the directivity of the transmitting antenna, and here two problems must be recognized. On the one hand, although the available effective power can be increased tremendously by the use of directive vehicle antennas, such antennas presuppose an ability to control the attitude of the space vehicle. On the other hand, it is generally desirable to provide failure instrumentation for space vehicles so that, if an element of the system fails, it is possible to determine the cause of the failure. For this purpose, it is desirable to have an omnidirectional vehicle antenna. It is probable that future space vehicles will use omnidirectional antennas for minimum tracking and failure instrumentation, and directional antennas for the primary transmission of information.

An empirical relation was derived from present experience with high-frequency components which leads to the parametric relations shown in Fig. 1 between the gain, weight and frequency of parabolic antennas, and between the power, weight and frequency of the transmitter system including an omnidirectional solar cell power supply and temperature control equipment.

Essentially these curves are statements of the fact that the antenna gain is proportional to the square of the diameter in wavelengths and the weight of an antenna is proportional to its area, and that the weight of the power supply and of the heat sink increases with frequency since the efficiency of available transmitters decreases with increasing frequency.

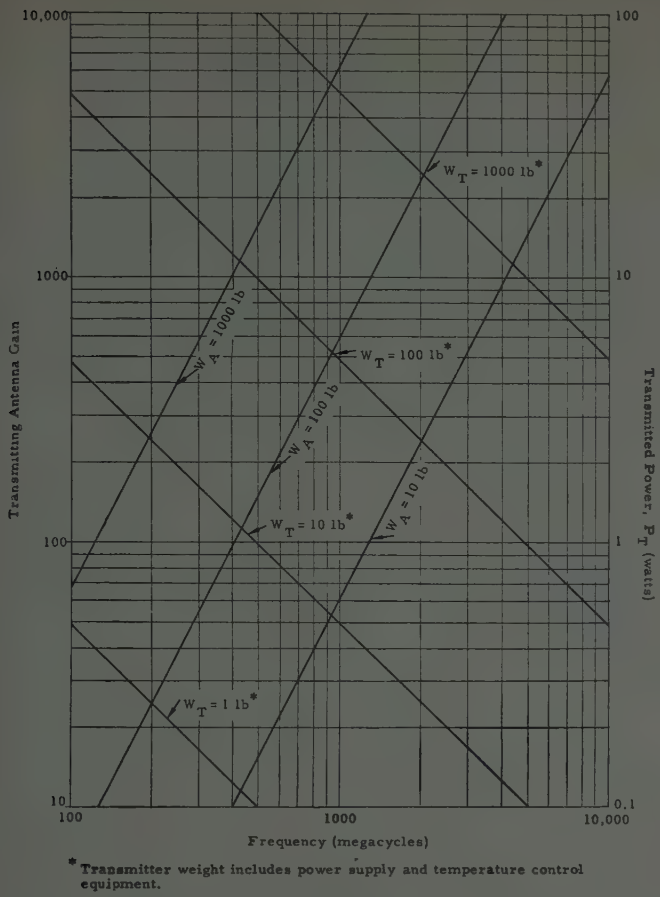


Fig. 1—Airborne antenna gain for three antenna weights, W_A, and transmitted power, P_T, for four transmitter weights, both as a function of frequency.

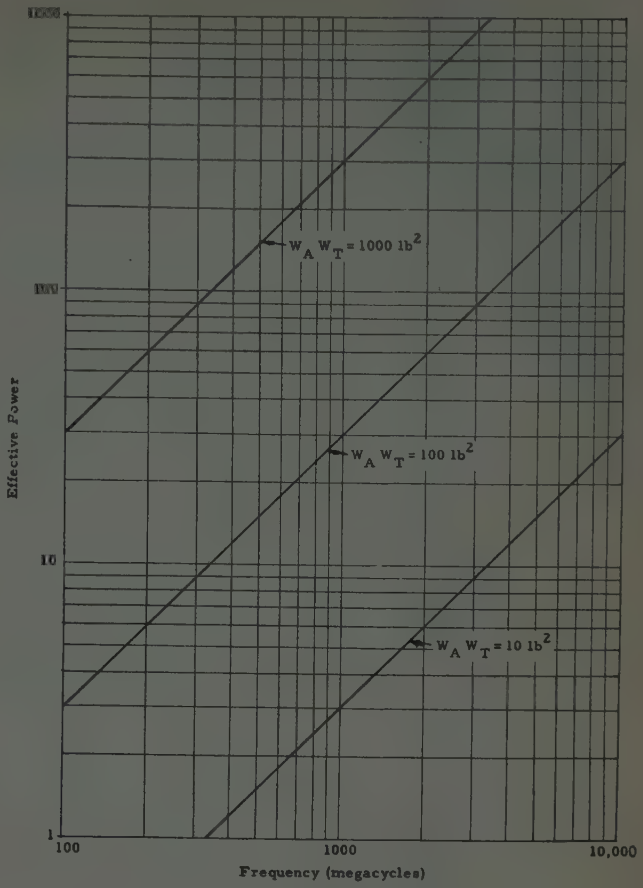


Fig. 2—Effective radiated power as a function of frequency for three given weights to be allocated to transmitting antennas and transmitters.

Multiplying the transmitted power by the antenna gain gives the effective radiated power, which is plotted as a function of frequency in Fig. 2. The parameter here is the product of the weight allocated to the antenna system and the weight of the transmitter and its power supply. The curves are derived from

$$\text{Effective Power} = P_T\, G \approx 0.3\, \frac{W_A W_T}{\lambda} \tag{1}$$

where

P_T = the transmitted power,
G = the antenna gain,
W_A = the weight of the antenna,
W_T = the weight of the transmitter and power supply, and
λ = the wavelength in feet.

Regardless of the constant, which varies from system to system, it is apparent from the form of the equation that an optimum division of weight would make the antenna system weight equal to that of the transmitter system. It is also apparent where directive antennas are used that, over this range of frequencies, the higher the frequency the greater the effective transmitted power. It must be noted, however, that this simplified analysis has not taken into account the increased problems associated with attitude control as the antenna bandwidth narrows, nor the increased mechanical tolerance at high frequencies. These are important as one approaches 10,000 mc, and, at such frequencies, the curves are no longer valid.

In the case of failure instrumentation where a nondirective satellite antenna is desirable, the curves of transmitter power vs frequency of Fig. 1 apply, and it is evident that the lower the transmitting frequency the lighter the system is. However, as is shown in the next section, galactic noise limits the lowest permissible frequency for space communication to the order of 400 mc.

The Ground Receiving System

The receiving system may be improved either by increasing the receiving antenna area or by decreasing the effective noise temperatures in the input circuits of the receiver. If the receiving system is mounted on the earth, then even with the best of maser preamplifiers, the input noise temperature is limited to 50°K due to the antenna radiation characteristics (see Fig. 3). Therefore, for ground-based antennas, no real improvement in input noise appears practical.

However, if the receiving antenna is in space, then it is possible to achieve an input noise temperature which is limited only by the galactic background noise and the inherent characteristic of either maser or parametric amplifiers. With such antennas, it appears that an effective noise temperature on the order of 1°K is possible and that as techniques for the maser preamplifiers are developed even lower temperatures may be achieved. The variation of these limitations with frequencies are shown in Fig. 4.

The remaining system parameter which may be varied is the effective area of the receiving antenna. The following paragraphs show that the steerable parabolic reflecting antenna is the most practical design for ground-based antennas. This type of antenna has been built in sizes up to 250 feet in diameter, and studies have been made indicating that sizes up to 600 feet are feasible. However, for diameters larger than about 250 feet, it is necessary to introduce an elaborate servocontrol system for adjusting the individual subpanels as a functioy of the antenna elevation angle and wind loading. Thus, there is a point at about 250 feet beyond which there is a relatively sharp increase in mechanical complexity and cost.

To investigate the effect of increasing antenna size, one can employ the usual formula for power received as a function of power transmitted. This equation is

$$P_R = kTBn\left(\frac{S}{N}\right) = \frac{(P_T G_T)A_R \epsilon}{4\pi R^2 l} \qquad (2)$$

where

$B = 10$ cps = bandwidth,
$n = 2.0$ db = noise figure of receiver,
$S/N = 10$ db = required signal-to-noise ratio,
$\epsilon = \approx 0.5$ = antenna efficiency,
$kT = 1.3 \times 10^{-23} \times 300 \cong 4 \times 10^{-21}$,
G_T = vehicle antenna gain,
P_T = vehicle transmitter power,
$R = 50{,}000{,}000$ miles $= 3 \times 10^{11}$ feet = range,
$A_R = \pi D^2/4$ = antenna collection area (D diameter), and
$l = 6$-db loss from polarization, cables, etc.

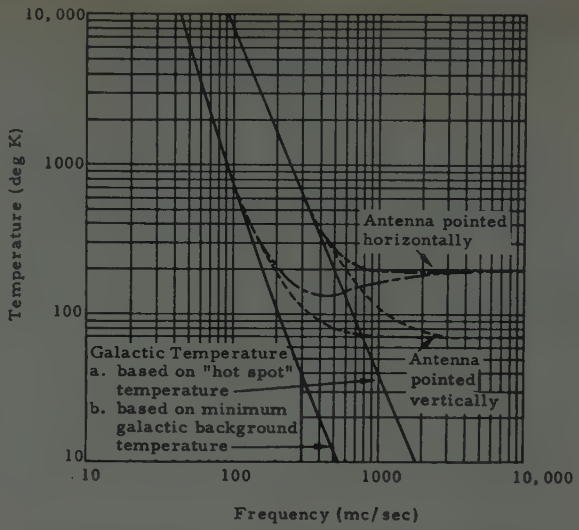


Fig. 3—Equivalent antenna temperatures including galactic background, atmospheric attenuation, and antenna sidelobes. (Assumes an aperture efficiency of 0.5, that half the sidelobes' energy is received from the ground, which is assumed to be a blackbody at 280°K.)

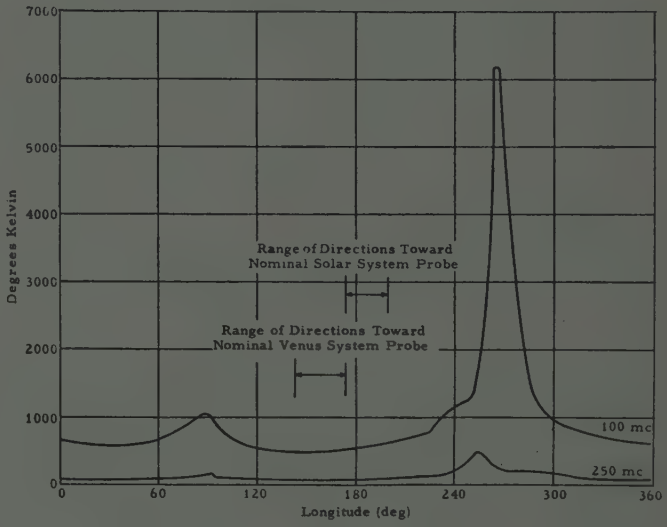


Fig. 4—Cosmic radio noise in degrees Kelvin in the ecliptic as a function of longitude (measured east from the vernal equinox).

The figures chosen are representative of those attainable in extreme range space communication systems. One will note that the bandwidth, 10 cps, is relatively narrow compared with that of the usual earth-type communication systems. The noise figure of the receiver is representative of current good practice in receivers in the 100- to 400-mc range. This figure does not represent noise figures attainable using parametric or maser amplifiers which are discussed later. A signal-to-noise ratio of 10 db is quite low, since this signal-to-noise ratio must be adequate for acquiring after search as well as for tracking once the signal has been found; in fact, a 20-db signal-to-noise ratio would be desirable. The ground antenna efficiency of 0.5 leaves some margin for sidelobe control. Vehicle antenna gain and vehicle transmitter power are treated, as they in fact are, as functions of each other (see Fig. 2). The 6-db losses are conservative, but reasonable, assumptions.

The range chosen, 50,000,000 miles, is not the range to any particular point in space, but it is representative of ranges from the earth to interplanetary vehicles on their outward journey. It does not represent, for instance, the maximum that one might encounter if an attempt were made to follow the vehicle over a complete circuit about the sun.

Solving (1) then for D, the antenna diameter, one obtains

$$D = 4R\left[\frac{lkTBn(S/N)}{(P_T G_T)\epsilon}\right]^{1/2}. \tag{3}$$

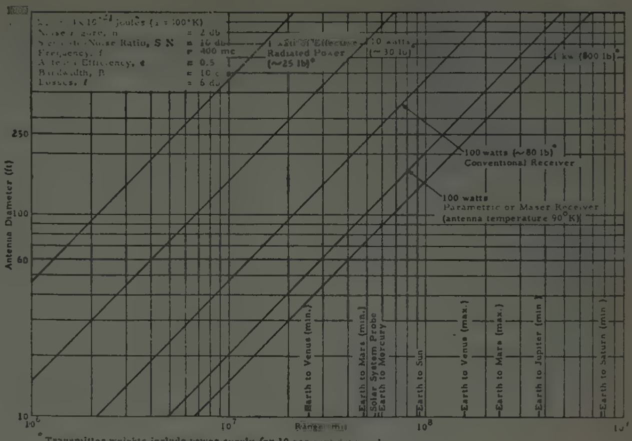
One interesting fact, apparent at once, is that the antenna diameter required is directly proportional to range. Using the numbers assumed above in (2), one obtains a required antenna diameter of about 250 feet. It is obvious from the square-root relationship that any reasonable variation in all parameters except range will have only a moderate effect upon the required antenna diameter.

As another illustration of this relationship, the JPL Microlock receiver and a 250-foot ground antenna would receive a usable signal of 10-cycle bandwidth with an effective radiated power of 100 watts in orbit about Venus at about 55,000,000 miles. The transmitter would weigh about 15 pounds. Optimum solar cells and batteries would weigh about 50 pounds, assuming a 10 per cent duty cycle. To use a 60-foot antenna would require either 1000 watts of effective radiated power or a parametric or a maser preamplifier with a 200-watt transmitter, or the bandwidth would have to be reduced to 0.5 cps.

Fig. 5 illustrates the effect of range upon the ground antenna size for various effective radiated powers from the vehicle. This figure assumes a bandwidth of 10 cps, a signal-to-noise ratio of 10 db, a receiver noise figure of 2 db, and losses of 6 db. It shows that at 55,000,000 miles 100 watts of radiated power with a 250-foot antenna is just satisfactory.

Fig. 6 shows the required antenna diameter as a function of the range to the vehicle with three different ground receivers. A 100-watt transmitter with a conventional ground receiver can give a 10-db signal-to-noise ratio at 55,000,000 miles, using a 250-foot antenna. However, if a parametric or maser receiver (with an antenna temperature of 70°K) is used, a 100-watt transmitter can provide a usable signal at 120,000,000 miles. If the frequency is increased to 1100 mc, the range can be increased to 135,000,000 miles. Although parametric amplifiers and maser noise temperatures of less than 30°K are possible, the limiting noise will be determined principally by antenna temperature, as has been shown in Fig. 3. Fig. 6 again illustrates the principle that a great deal can be gained by improving the ground system—gains which become a permanent part of the system.

A more direct improvement is the use of variable receiver polarization. This is dependent on a knowledge or measurement of the plane of polarization of the received signal, but with relatively simple equipment, it will produce a 3-db increase in signal-to-noise ratio.



* Transmitter weights include power supply for 10 per cent duty cycle.

Fig. 5—Antenna diameter as a function of range for a number of transmitters.

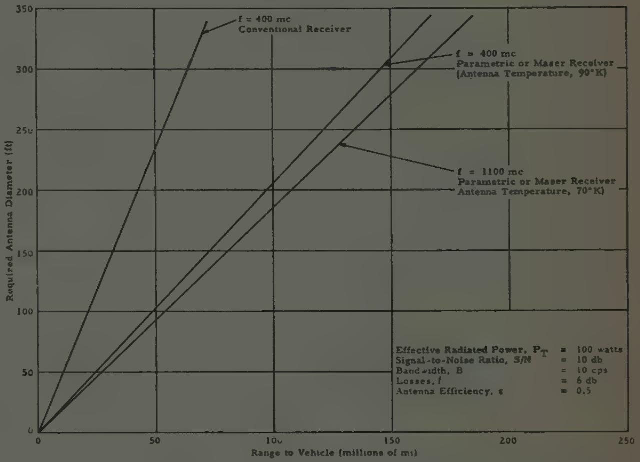


Fig. 6—Antenna diameter as a function of range for a conventional receiver and a maser receiver at two frequencies.

Figs. 7–9 are cross plots of Fig. 5 (with different parameters varied). Fig. 7 shows the signal-to-noise ratio as a function of range of the vehicle for a number of antenna sizes. Fig. 8 shows the bandwidth as a function of range to the vehicle for a number of antenna sizes. Fig. 9 shows transmitted power requirements as a function of range to the vehicle for a number of antenna sizes.

Each of these figures shows the great gains that are made by increasing the antenna diamter. In particular, Fig. 9 shows that almost 1600 watts of radiated power are required to transmit 55,000,000 miles from the vehicle to a 60-foot ground antenna with a signal-to-noise ratio of 10 db. If a 250-foot antenna is used, only 100 watts are required.

In connection with Fig. 7, it should again be emphasized that a signal-to-noise ratio of less than 10 db is not satisfactory since that ratio or more is desirable for acquisition.

The parameters used above are not all representative

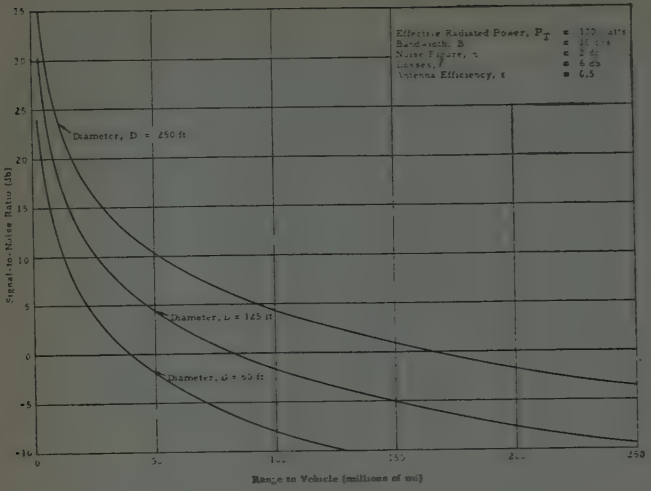


Fig. 7—Signal-to-noise ratio as a function of range for three ground antenna sizes.

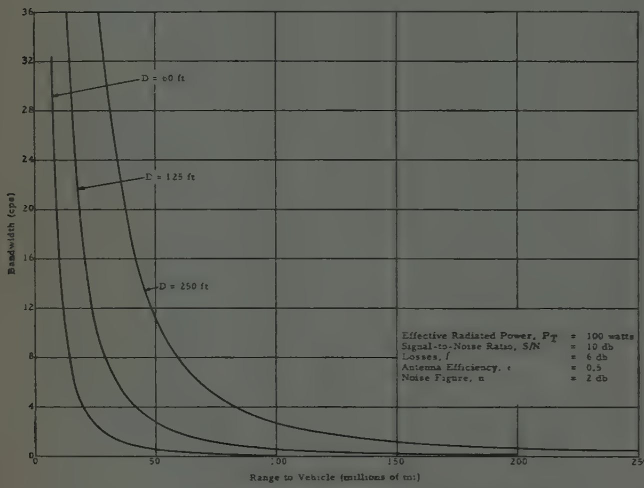


Fig. 8—Bandwidth as a function of range for three ground antenna sizes.

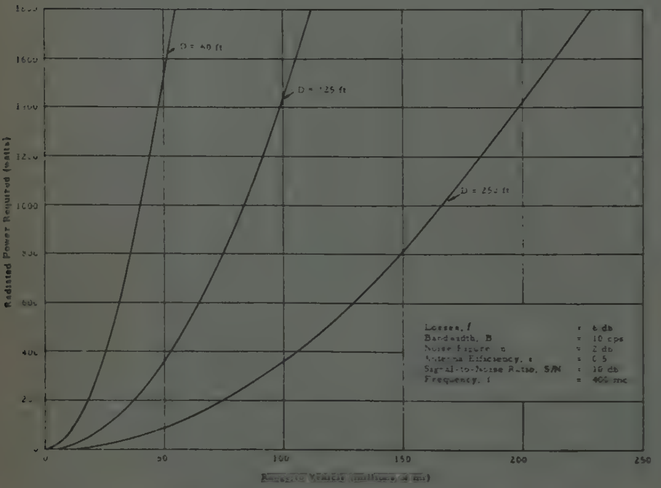


Fig. 9—Airborne transmitted power as a function of range for three ground antenna sizes.

of what one might ultimately desire in a space communication system. For instance, it seems reasonable that we would ultimately wish to transmit facsimile pictures with bandwidths of hundreds of thousands of cycles or possibly even TV pictures with bandwidths which exceed a megacycle. Manned space flights will require, as an absolute minimum, a continuous voice channel of at least telephone quality. For a continuous voice-communication capability independent of vehicle attitude, 500 watts of continuous power drain, which is near the optimum for 100 watts of effective radiated power from a communication transmitter at 400 mc, is necessary. This is a considerable drain on a space power supply. It is apparent that to achieve large bandwidths at long distances requires, in addition to the use of large ground antennas, improvement in all elements of the communication system.

To determine the type of ground antenna, the approximate operating frequency must first be established. An examination of presently available transmitters, of the uncertainty concerning the operating frequency of low-noise maser amplifiers, and of the rate of development of efficient microwave transmitting tubes, clearly underscores the necessity for an antenna capable of operating over a wide range of frequencies. Since multielement antenna arrays with a large effective collecting area are essentially fixed frequency devices, it appears that the frequency insensitive reflector antenna is better. Moreover, since the space vehicles will have to be tracked over large angles and since antenna arrays are essentially fixed in direction, the steerable parabolic dish is the only choice which satisfies all the requirements.

As the physical size is increased, the efficiency of the capture area must be maintained. However, to maintain area efficiency, the physical tolerances must be maintained to about $\frac{1}{16}$ of a wavelength to focus the received energy into the feed. Thus, if the wavelength is one foot and the antenna is 500 feet in diameter, the tolerances would be about one part in 10^4, or a $\frac{1}{2}$-inch total deflection. Despite these complex engineering problems, very large antennas have been built (Fig. 10 shows the 250-foot radio telescope near Manchester, England) and many more are contemplated.

Given only initial specifications as to sky coverage, antenna size, shape, and accuracy, every group that has studied large antennas has come up with essentially the same design; that is, a merry-go-round or azimuth tracking stand upon which is mounted an antenna on a horizontal axis so that the elevation angle can be varied. This is substantially the design evolved at Manchester for the Jodrell Bank antenna.

The only existing antenna suitable for interplanetary space communications, the 250-foot British antenna at Jodrell Bank, England, is used in the only world-wide tracking network in the world today, the space navigation network (SpaN-Net) of STL/BMD. This antenna has demonstrated, during the Able[1] flights, the value of

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[1] Pioneer I and II, Explorer VI.

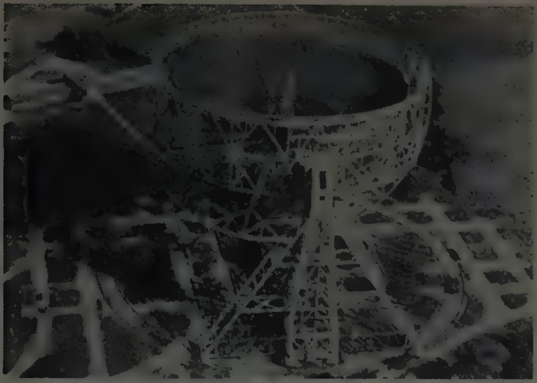

Fig. 10—The radio telescope near Manchester, England.

large antennas. However, because it is located about 55 degrees north of the equator, its coverage of the plane of the ecliptic is not optimum. In addition, since it is the only existing at present, the coverage of any given point in space is limited by the rotation of the earth.

INFORMATION ENCODING AND TRANSMISSION

In a communication system, the kind and amount of information to be transmitted is of first importance. The kind of information depends in large measure upon the source and, as the following examples indicate, it may be varied by proper selection of the form of the message generated. The amount of information must be tailored to the channel capacity of the communications system and, for many scientific purposes, may be varied without much loss in the utility of the experiment. For instance, the size of the diaphragm on an impact-sensing micrometeorite detector may be reduced until the number of impulses per second is adjusted to the available bandwidth. Or, in the case of a scintillation or Geiger counter, appropriate scalers may be inserted so that the final output counting rate is a small, but known, fraction of the input rate. This type of system has, in addition, the possibility of easily varying the output rate by switching the number of divider stages. In case there is a requirement for transmitting pictures, the effective bandwidth and information rate may be reduced by sampling techniques.

Another kind of information is obtained from the various equipment performance monitors in the vehicle. Since the information rate from these monitors is usually very low, it is possible by subcommutating to transmit a very large number of measurements over a very narrow bandwidth. Thus, it is possible by relatively simple means to adjust the information rate for physical measurements to the available channel capacity of a space communication system. However, for such information as real time voice channels there is only a limited possibility of reducing the bandwidth even though the information rate for this service is small, since there is a real limitation on the efficiency of encoding represented by our vocal cords. Since voice or video transmission characteristics are not easily modified, and since the weight limitation on equipment is likely to persist for some time to come in space vehicles, the probability of achieving real time voice or video communications at interplanetary distances is rather low.

Basically, space communication suffers from trying to achieve two mutually incompatible objectives. On the one hand, there is the desire to transmit as much information as possible. On the other hand, there is the necessity to conserve weight with the resultant fundamental limitation on transmission time and information rate. As was shown in Fig. 9, the bandwidth is extremely limited even at fairly short interplanetary distances and for very large ground antennas. This limitation upon bandwidth sets the channel capacity, and thus the information rate which can be transmitted. Fig. 9 also demonstrates that an essential characteristic of a space communication system should be flexibility in bandwidth, so that when the vehicle is at close ranges, its bandwidth can be large, while at long ranges its bandwidth may be narrowed.

The method of modulation establishes the efficiency of the communications system. As Hartley and Shannon[2] have shown, there is a maximum amount of information which can be transmitted through a channel of given bandwidth and signal-to-noise ratio. Thus

$$H = B_M \log_2\left(1 + \frac{S}{N}\right) \tag{4}$$

where

H = the information rate in bits per second,
B_M = the bandwidth of the message channel,
S = the signal, and
N = the noise in the bandwidth B_M.

Furthermore, it is important to note that the effective signal-to-noise ratio in the message channel is a function of the ratio of the transmission channel bandwidth B_T, to the message channel bandwidth and in the ideal case approaches

$$\left(\frac{S}{N}\right) B_T = \left(\frac{S'}{N'}\right) B_M. \tag{5}$$

There is a wide variation in the ability of modulation systems to approach the ideal. Thus, in the conventional FM broadcast system,

$$\left(\frac{S}{N}\right) = \left(\frac{S'}{N'}\right) \frac{B_M}{B_T} \tag{6}$$

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[2] R. V. L. Hartley, "Transmission of information," *Bell Sys. Tech. J.*, vol. 7, p. 535; July, 1928; and C. E. Shannon, "A mathematical theory of communication," *Bell Sys. Tech. J.*, vol. 27, pp. 379–423, 623–656, July/October, 1948.

and although the noise improvement in the output is real, it falls far short of the Shannon limits.

A comparison of the various modulation methods results in the conclusion that for space communications the most efficient system is probably some form of pulse code modulation, both from the standpoint of encoding and decoding, and from the standpoint of modulation efficiency. For purposes of illustration two forms of modulation adapted to space communications will be described and compared.

In the Pioneer series of lunar satellites, an FM/PM analog telemetry system developed by Jet Propulsion Laboratories was used.[3] This "Microlock" system uses a phase-locked receiver to achieve a lock-on sensitivity of −150 dbm with a 10-cycle locked-loop bandwidth. There are six subcarriers with a theoretical modulation bandwidth of 0.8 cps, although in practice, because of subcarrier drift and the dynamic characteristics of the subcarrier discriminators, they have a realizable information bandwidth of only about 1/100 cps. As was demonstrated, this system has the necessary sensitivity to operate to lunar distances with a transmitted power of about 100 mw. The combined total information rate for the six channels is about the equivalent of 0.5 bit per second. The limiting characteristics of this system are:

1) the information bandwidth is fixed;
2) the practically achievable channel efficiency is low;
3) since the system is not quantized, it suffers degradation on retransmission or rerecording; and
4) the information must be transmitted in real time.

In an attempt to improve the system, a digital telemetry system called "Telebit" was developed which will, at lunar distances and the same 100-mw power, permit the transmission of 8 bits of information per second, or, on command and for transmission at greater or lesser distances, will change power and transmit either 1 bit or 64 bits per second. In addition, with airborne analog-to-digital converters, the information is quantized and digitalized, so that once a message is received, it will not be degraded by retransmission over communication links to the central station. The Telebit system provides, in addition, a transistor memory for storing the output of the experiments, so that intermittent transmission of the data is possible.

A quantitative comparison of the Pioneer I and II Microlock and the Telebit system with the ideal communications system described by Shannon[4] leads to the following expressions:

$$\beta = \left(\frac{S}{N}\right)\frac{B}{H} \tag{7}$$

where β is a figure of merit based on Shannon's theorem, (4); for an ideal system where the transmission channel bandwidth is infinite, the figure of merit

$$\beta_0 = 0.693;$$

for the case of the Microlock system used in Pioneer I and II

$$\beta = 43.5,$$

or

$$\beta/\beta_0 = 18 \text{ db};$$

while for the Telebit system

$$\beta = 4.5,$$

or

$$\beta/\beta_0 = 8 \text{ db}.$$

It is interesting to note that this latter figure is nearly as good as the best quantized system described by Saunders[5] for this bit level and appears to approach the best physically realizable modulation system.

Since the Telebit system has not been described elsewhere a brief description follows. Basically, Telebit is a digital telemetry system which provides for transmission of information from space probes to the ground and presents the information in a partially processed form. The payload portion of the Telebit system accepts information from both analog and digital experiments and processes this information until it is in a form suitable for transmission. The ground portion of the Telebit system consists of several stations located throughout the world and a central processing station for rapid processing and presentation. The payload portion of the Telebit system permits accumulation of information during periods between transmissions. It also commutates several types of input information so as to produce a time-multiplexed train of pulses containing information about a variety of experiments.

The Telebit system accepts two kinds of inputs from the experiments. The first consists of information in the form of the occurrence of an event or events, such as micrometeorites and radiation particle impingement. The second class consists of analog inputs where the variable to be measured is a continuous function of time.

Power considerations set bounds on the time during which information can be transmitted to the ground and upon the period between such transmissions during which batteries are recharged. Sufficient time exists during each allowable transmission period to permit a readout of each primary experiment two or three times. The Telebit system has been designed with the intent of optimizing information accumulated during transmitter OFF times while permitting maximum information transmission during the brief ON times. A careful review of high-energy particle experiments, for instance, indicated that 10 primary digits would adequately accumulate the number of events during a transmitter ON time.

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[3] "A Coherent Minimum Power Lunar Probe Telemetry System," Jet Propulsion Labs., Calif. Inst. Tech., Pasadena, Calif., external publication No. 610; August 12, 1959.

[4] Shannon, *op. cit.*, p. 654.

[5] R. W. Sanders, "Digilock Telemetry System," 1959 National Symposium on Telemetering, San Francisco, Calif., September 28–30, 1959.

Analog information does not directly lend itself to any simple means of accumulation. The experimenter must thus be content with obtaining analog information only during transmitter ON periods. Provision has been made, however, in some experiments, to provide for changes of scale as a function of the quantity being measured. An analysis of the accuracy with which analog information is obtained indicates that most primary experiments could be adequately described by six bits while some secondary experiments can be characterized by four bits. By grouping a six-bit experiment with a four-bit experiment, a word of 10 bits is derived. Conversion of the analog data to digital form is accomplished in an analog-to-digital converter which makes use of a 64-level digital ramp.

In addition to its function of accumulation and analog-to-digital conversion, the Telebit system commutates the successive experiments and thus derives a sequence of pulses which in groups characterize these experiments. The information conveyed during each commutated segment is called a "word" and the sequence of all words is called a "frame." While the number of pulses which comprise a word could take any value, a Telebit word is composed of 10 information pulses. Similarly, the number of words which compose a frame is determined by the number of experiments. The Telebit system includes 10 information words.

In addition to the information pulses of each word and the information words of each frame, synchronizing symbols are inserted to ease the decommutation problem on the ground. Two synchronizing pulses always having the same form are added to each word and one synchronizing word is added to the set of each frame. Thus, an entire word consists of 12 pulses and an entire frame of 11 words.

Inasmuch as it is desirable to vary the information transmission rate as the range to the payload changes, three pulse rates were provided, 1, 8, or 64 per second. For these rates it takes approximately 132 seconds, 17 seconds, or 2 seconds to transmit one frame.

One of the analog words has a 16-element subcommutator associated with it. This subcommutator has the capability of selecting in sequence any one of 16 slowly varying quantities. Thus, at the three transmission rates it will take approximately $\frac{1}{2}$, $4\frac{1}{2}$, or 35 minutes to complete an entire subcommutator cycle.

The method of transmitting the telemetry information to the ground is to biphase modulate a subcarrier which in turn phase modulates a radio-frequency carrier. This technique has the advantages of providing a continuous carrier for acquisition and tracking from the ground and an unambiguous resolution of the pulse data through the use of an adjacent pulse comparison biphase demodulator without a ground coherent oscillator. A disadvantage of this technique is an approximately 6-db loss in signal-to-noise ratio because of the limited subcarrier modulation power.

Areas of improvement are the use of biphase modulation of the carrier, the use of multiple bit encoding, and the sophistication of the logical design of the system.

In addition the Telebit system may be improved by expanding analysis or processing of the experimental data before transmission. For example, in the present Telebit system, the micrometeorite count is stored for a period of 6 to 10 hours and thus the resulting transmission from the satellite to the ground provides an average rate for a 6-hour period and then, as the experiment is read out again, the average rate for a $2\frac{1}{2}$-minute interval. A relatively simple change in the digital circuit logic would permit one to record during the 6-hour period the maximum counting rate that occurred in any one of the $2\frac{1}{2}$-minute intervals, the minimum counting rate that occurred in any of the $2\frac{1}{2}$-minute intervals, and the times at which these maximum and minimum rates occurred, in addition to the average count for the entire six-hour period. By the skillful application of such satellite data processing, considerable increases in the amount of information obtained from the experiments can be achieved without appreciably increasing the requirements on the total information channel capacity.

Finally, as data is accumulated, experimenters will be better able to define the steps required for the analysis of the data. At this point it will be possible to program the central computer to accomplish the mechanical part of the data analysis in addition to its normal task of data reduction. This ability to approach the ideal of real time analysis of experimental data is perhaps the most important single characteristic of the digital telemetry system.

THE PRESENT BOUNDARIES OF SPACE COMMUNICATION

The preceding paragraphs have developed the limitations and the characteristic parameters of the transmitter, the receiver, and the modulation subsystems, which together make up the building blocks of any space communication system. These results may be combined and summarized as in Fig. 11, where the information rates that can be achieved with present, or at least reasonably foreseeable, space communication systems are plotted against range to the vehicle. The curves assume that an optimum distribution of the weight between the satellite antenna and the satellite transmitter is achieved and that the best parametric amplifiers are used in the receiver. Several combinations of receiver-antenna area and transmission system weight are assumed which are characteristic of the several space missions.

From these curves it is evident that for low-altitude satellites, and reasonable payload weights, very large amounts of information can be transmitted. For instance, with ten pounds devoted to the transmitter and antenna system, it is possible to send a television picture from a satellite in a five-hundred mile orbit to a 60-foot antenna on the ground. Referring back to Fig. 1,

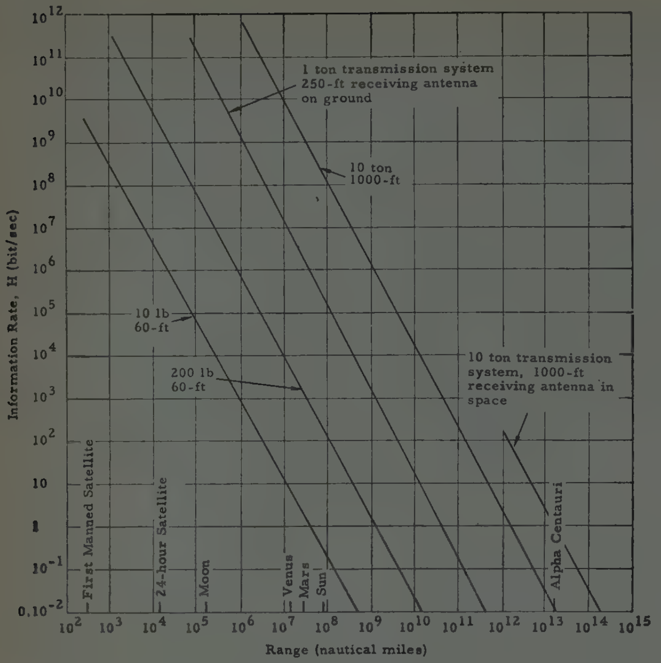

Fig. 11—Information rate, H, as a function of range for various combinations of payload antenna and transmitter weights and receiving antennas.

this corresponds to a transmitter power of only $\frac{1}{10}$ of a watt and an antenna gain of about 100.

As another example, in a 24-hour orbit it is possible, with 200 pounds of transmitter, power supply and antenna, to send 10^9 bits per second to a 60-foot antenna on the earth, which is the information contained in about 200 television channels.

Another class of space mission is that involving exploration of the moon. For this mission, the same 200-pound system is capable of transmitting a television picture, or 10^7 bits per second of information from the moon to the earth. If only ten pounds can be devoted to the transmitter-antenna combination, the maximum amount of information drops to 10^4 bits per second, and if a nondirectional antenna is necessary, the maximum information drops to 100 bits per second for a ten-pound weight.

The third class of space mission is the exploration of Venus or Mars, an era of space communication which is almost upon us. Here the ten-pound system, with an optimum antenna-transmitter, is capable of transmitting only one bit per second, although the 200-pound system will provide an information rate on the order of 1000–5000 bits per second at the minimum distance between earth and Mars or Venus. If transmission is desired from a satellite circling Mars or Venus, then the 200-pound transmitter will permit transmission of 100 bits per second over the entire path of Mars, providing a 250-foot diameter antenna is used on the earth. To transmit video from a satellite orbiting Venus or Mars will require transmission systems of between one and ten tons, and ground antennas between 250 and 1000 feet in diameter. As an example, 50 million bits per second can be transmitted from Mars in its extreme position using a ten-ton optimum payload with a 1000-foot receiving antenna on the ground.

Perhaps the most challenging space mission involves communication with an interstellar probe. Even with a ten-ton transmission system and a 1000-foot receiving antenna on the earth's surface, it is barely possible to track a probe to the nearest star, Alpha Centauri. The information rate at these distances is so low (only 1/100 bit per second), and the required oscillator stability so high (one part in 10^{12}), that such a mission is not, strictly speaking, within the present state of the art of our components. One alternative that provides reasonable, although still very small, information rates is to place the receiving antenna in a satellite orbit around the earth at an altitude such that the heat contribution of the earth's surface to the received energy in the antenna is small. If this can be done, the inherent low-noise properties of maser amplifiers may be fully realized. Then the ten-ton optimum system transmitting into a 1000-foot antenna at an altitude of about five thousand miles above the earth's surface would provide an information rate of about one bit per second from Alpha Centauri.

This last example is illustrative of the tremendous problems associated with direct communication over interstellar distances. If the problems of locating and stabilizing communication repeater stations could be solved, the proper placement of a hundred such vehicles would permit transmission of 100 bits per second to Alpha Centauri. Since the time of flight for these missions will be tens of years, a new order of component reliability is also required.

It is clear that both a considerable improvement in the entire transmission system and in our basic communication system concepts will be required before practical interstellar communication is feasible.

To summarize, the present practices in transmitter design, in solar cell power sources, in receiver design, and in antenna design, have been explored. The fundamental limitations on modulation efficiency have been examined, and two examples of modern practice have been described. Because of the limitations imposed by the conservation of energy, both in generating power and in the modulation of this power for the transmission of information, improvements in subsystems, although they will occur, are not likely to be order-of-magnitude improvements. Nevertheless, it appears that communication with large information rates is possible throughout the solar system.

ACKNOWLEDGMENT

The system studies considered here are in large part the result of creative effort of many of my colleagues. The paper itself is due in large measure to R. A. Park.

Propagation and Communications Problems in Space*

JOSEPH H. VOGELMAN†, FELLOW, IRE

Summary—**The problems of propagation and communications arising from landings on the Moon, Venus, or Mars are treated in terms of the characteristics required for the communication system to achieve data transfer between parties on the surface of these bodies and the communications problems arising in the transfer of information from these bodies back to the Earth. Consideration is given to Doppler shift, Faraday rotation, tracking and stabilization of antennas, and ground network requirements. The problems of communications between vehicles in space in terms of signal acquisition and antenna orientation and tracking are described.**

INTRODUCTION

THE recent literature has concerned itself with the problems of communications by means of vehicles in space serving as passive or active reflectors, and with problems of communications to vehicles traveling in space. This paper will treat the problems of propagation and communications arising from landings on the Moon, Venus, or Mars. Each of these will be treated in terms of the characteristics required for the communication system to achieve data transfer between parties on the surface of these bodies and the communications problems arising in the transfer of information from these bodies back to the Earth. The second case which will be considered is communications between vehicles in space.

GENERAL FACTORS

The frequency bands available for communications from the Earth to any other celestial body are limited by the radio frequency characteristics of the Earth's atmosphere. Accordingly, for communications to the Moon, Mars, or Venus, communications bands are available in Table I, where the frequency band from 100 to 10,000 mc appears to be the logical band of operation. Fig. 1 gives the transmission power in dbw/cps vs range for inverse square spreading assuming 0 db noise figure at the receiver. Fig. 2 gives the beamwidth in degrees at the half power points for parabolic antennas as a function of the antenna diameter. It also gives the effective gain achieved by these antennas. Fig. 3 shows the range which may be achieved at 50 mc. These curves use a receiver noise figure of 3 and a transmission antenna gain of 10^4. Values are given for combined transmitter and receiver antenna gains of 10^4 and 10^6 for receiver noise powers corresponding to values from 10^{-13} watts to 10^{-17} watts.

The obstacles to communications from the Earth to Mars, Venus, and the Moon are the following:

1) Doppler shift due to the relative motion of the body with respect to the Earth. These values are of the order of 0 to 106.4 cycles per megacycle of RF signal for transmissions from Earth to Mars and 0 to 112.7 cycles per megacycle for transmissions from Earth to Venus.

2) Doppler shifts due to the rotation of the Earth on its axis as well as the rotation of the other planets on their axes. The Doppler shift due to the rotation of the Earth for communication to the Moon varies from 0 to 0.199 cycle per megacycle of RF at the pole to 1.362 cycles per megacycle at New York City.

3) Faraday rotation due to the variable density of the Earth's atmosphere and its magnetic field. This rotation is maximum at the lowest frequency and a minimum at the highest. It is a maximum at the horizon and is essentially zero along a path perpendicular to the surface of the earth.

4) Tracking and stabilization of large antennas, possibly as large as 1000 feet on a side, are required to insure that the receiving station on the Moon, Mars, or Venus is within the main lobe of the antenna pattern.

5) Axial rotation of Venus and Mars will require a ground network on these planets to insure that the earth transmitters and receivers are in view.

6) In the case of the Moon one side is always in view of the Earth while the other is always invisible. This necessitates a communications system along the surface of the Moon or an appropriate lunar satellite in the event that communication with the far side is required.

The Doppler requirements discussed here would necessitate Doppler correction provisions or increased receiver bandwidths to insure that signals transmitted from the Earth would always be received in space. For an Earth to Mars link at 2000 megacycles this bandwidth would have to exceed 200 kilocycles even when the Doppler shift due to the rotation of Mars is neglected. This increase in diameter translates itself into the requirements for 13 db increase in transmitter power for a 10-kc information channel.

COMMUNICATIONS ON THE MOON

Line-of sight radio waves for transmissions along the surface of the Moon are of very short range due to the small radius of curvature of the Moon. Unless high mountain peaks can be used as relay points, this type of transmission does not appear desirable.

Low frequency ground waves of narrow bandwidth, however, appear entirely practical. The significantly lower gravitational field and the absence of wind and weather make extremely efficient large antennas practical at low frequencies.

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TABLE I

Frequency Band	Type	Characteristics
100 to 10,000 mc	Tropospheric	Scattered in troposphere Diffracted by obstacles Line of sight
10^4 to 7.5×10^8 mc	Millimeter and infrared	Line of sight Absorbed by atmosphere
7.5×10^8 to 1.5×10 mc	Light	Visible to human eye Scattered by atmosphere Absorbed by clouds

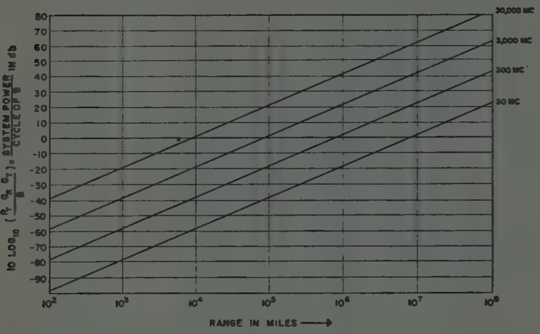


Fig. 1—System transmission power in dbw/cps vs range for inverse square spreading; 0-db noise figure.

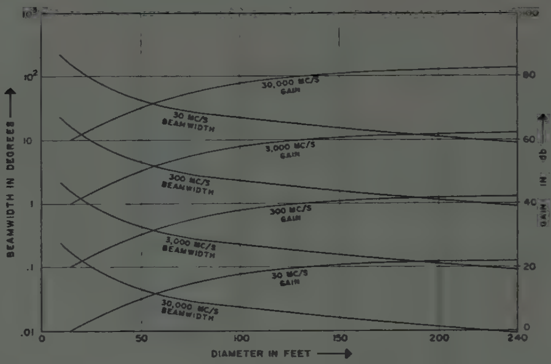
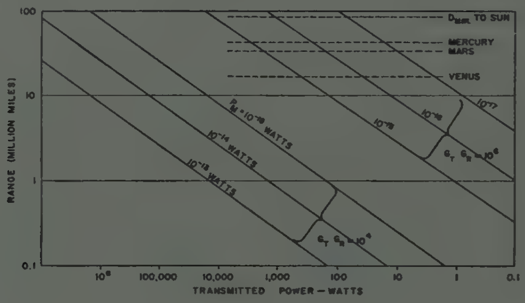


Fig. 2—Half-power beamwidth for parabolic antenna vs diameter



For other frequencies multiply transmitter power by

$$\left(\frac{fmc}{50}\right)^2$$

$$R = \frac{1.48\times10^2}{fmc}\sqrt{\frac{P_T G_T G_R}{P_M}}$$

$P_M = KTB\,N_f$ (losses) (integ. loss)
$K = 1.380\times10^{-23}$ joules/°K
T = temperature—°K
B = bandwidth—cps
N_f = noise figure

Fig. 3—Range of one-way radio at 50 mc.

Alternately, the lunar relay satellite of the active type may be used to provide the necessary path to a retransmitting station on the near side of the Moon.

Communications on Mars or Venus

A significant increase is required in information regarding the atmospheres of Mars and Venus, the nature of their magnetic fields and the extent of the ionized layers above their surfaces. Until such information becomes available, it must be assumed that surface communications will be limited to line-of-sight waves and the possible employment of a satellite orbiting these planets. The possibility of using ionospheric reflecting modes, tropospheric and ionospheric scatter as well as meteoric scatter cannot be determined at this time. Atmospheric conditions on these planets may impose frequency restrictions considerably different from those encountered on the Earth. It is to be hoped that the radio frequency windows on these planets are compatible with the radio frequency windows on Earth.

The major restriction on communications from the Earth to Mars and Venus is the presence of the sun along the radio path between the Earth and these planets for a portion of their orbits. The sun is a radio frequency source of tremendous power and is considerably closer to the Earth for a good portion of the planetary orbit. As a result, it will prevent communication for a considerable portion of the time. This radio silence can only be overcome by means of a solar satellite having an orbital path such that it permits both the Earth and the planet to view the repeater satellite without receiving the radio energy of the sun. The satellite will also require directive antennas to minimize solar radio noise in its receiver.

Communications Between Vehicles and Space

The vacuum that is space imposes no frequency restrictions on the communications devices. Redirected modulated sunlight or solar radio energy can be used as well as radio waves. The major problem of vehicle to vehicle communication is the requirement that each vehicle should know where it is located with respect to the other. The acquisition problem is extremely difficult. The possibilities for acquisition are as follows:

1) A preplanned communications program for the two vehicles would require that the orbiting information of each vehicle be precisely known to the other and that computational equipment be included in the vehicle for the orientation of antennas.

2) Use of the Earth as a relay is another possibility. With the Earth maintaining continuous track of all vehicles in space, it would be possible for a vehicle to communicate to the Earth and have the message repeated and redirected to the second vehicle.

3) Alternately, the Earth may be used as an acquisition director furnishing continuously or "on request" position information for all vehicles in space.

Once acquisition has been achieved, tracking in space is possible by using a continuous reference transmission for receiver antenna lock-on and by synchronizing the transmitting antenna to the receiving antennas.

Conclusion

Table II summarizes the problems of communications in space between vehicles, and between the Earth, Mars, and Venus.

TABLE II

Location	Frequency Band	Obstacles
Earth to planet	100 to 10,000 mc	Doppler shift Faraday rotation Tracking and stabilization
Space to space	3 mc	Antenna size
	3 to 10,000 mc	Antenna size Power
	10^4 to 10^{10} mc	Tracking and stabilization
On the moon	3 mc (ground wave)	Antenna size Power
	3 to 10^{10} mc (line of sight)	Very short range
On the planets	3 mc (ground wave)	Antenna size
	3 to 30 mc (reflected wave)	Ionosphere ?
	30 to 10,000 mc	Scatter ? Meteors ? ?
	10,000 mc	?

Propagation-Doppler Effects in Space Communications*

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Summary—**A generalized form of the Doppler equation is derived by field theory. The equation permits consideration of the effect of propagation phenomena on the Doppler shift under space flight conditions. As an example of an application, the effect of a medium which is nonhomogeneous in the flight direction is investigated. A deviation of the longitudinal and a transverse Doppler shift are the consequences. The magnitudes of both effects, which have to be taken into account in tracking, navigation, and Doppler evaluations, are considered.**

INTRODUCTION

DUE to the high velocities encountered in space flight, the Doppler phenomenon becomes very pronounced. It causes undesired frequency deviations which affect the operation of any kind of communications system, and which have to be taken into account, but it can also be utilized as a most useful tool in the physics and technology of space exploration, navigation, and tracking.

It can be anticipated that space flight conditions will affect the Doppler shift and that effects heretofore neglected will become apparent.[1] Wave propagation phenomena are among these effects. The anticipated influences were in fact observed and discussed in connection with Doppler measurements of radiations transmitted by satellites.[2] Other effects are due to relativity but, at the present state of the art of rocketry, these are not of significant magnitude.

It is the purpose of this study to investigate the effects of propagation phenomena on the Doppler shift. It may be noted for clarification that the effects considered as propagation phenomena are those which are not directly apparent from the commonly used equation for the Doppler shift, $\Delta\omega = \omega(v/c)\cos\alpha$, where ω is the operational frequency, v and c are the velocities of the vehicle and of the waves, respectively, and α is the angle between the direction of wave propagation and the flight path. These effects are primarily due to the nonhomogeneities of the medium (ionosphere, electron clouds) along the flight path of the vehicle or along the propagation path of the waves between the vehicle and the other station.

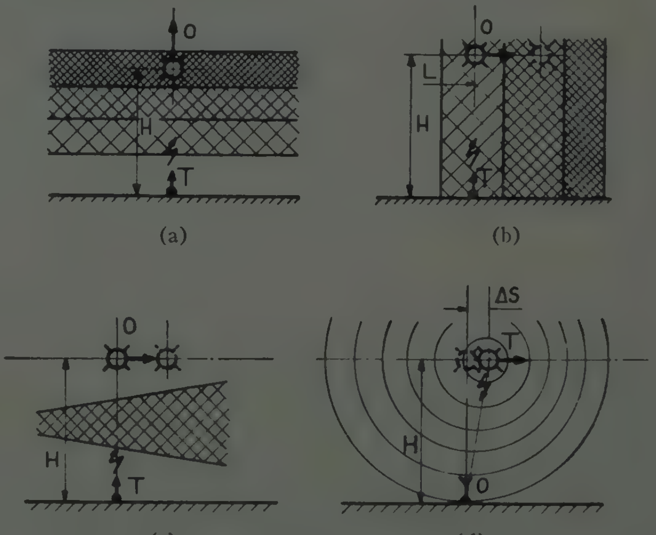
To study the effects of propagation phenomena on the Doppler shift, the above equation is not sufficient. A new formulation is necessary. Equations for the Doppler shift based on an optics method, which is an approximation, were derived previously.[1] The rigorous derivation must be based on field theory.

The first part of this study consists of a derivation of the Doppler shift by field theory, taking into account nonhomogeneities of the medium. The derived relations are used in the latter part to consider typical deviations from the regular Doppler shift due to the influence of the propagation phenomena.

Let us, as an introduction, consider some examples of propagation phenomena which affect the Doppler shift. Fig. 1 shows the operational conditions under which they occur. Fig. 1(a) indicates a vertical stratification of the medium, as it is typical for ionospheric layers. A transmitter T is placed on the ground, and the vehicle, in which an observer O is located, travels at a height H with a velocity v in the vertical direction. The question arises: How is the Doppler shift affected by the propagation phenomena?

The assumptions made for the second example are shown in Fig. 1(b). Such a stratification along the flight path occurs if the satellite travels in a region of varying electron density (border between day and night) in the horizontal direction. The resulting frequency deviation may be called a "transverse" Doppler shift. No Doppler shift would be observed without the influence of the medium.

A third kind of propagation effect is indicated in Fig. 1(c) with a layer of varying thickness, in which the propagation constant is different. An electron cloud of varying thickness may have a similar effect. Again, a transverse Doppler shift may be observed.



(a) (b)

(c) (d)

Fig. 1—Examples for the operational conditions for irregular Doppler shifts.

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* Original manuscript received by the IRE, December 2, 1959.

† The Ohio State University, Columbus, Ohio.

[1] F. J. Tischer, "Doppler phenomena in space communications," IRE TRANS. ON COMMUNICATIONS SYSTEMS, vol. CS-7, pp. 25–30; May, 1959.

[2] P. R. Arendt, "Preliminary results of measurements on Doppler shift of satellite emissions," IRE TRANS. ON ANTENNAS AND PROPAGATION, vol. AP-7, pp. 99–101; January, 1959.

Concerning the transverse Doppler shift, it may be noted that the finite travel time of the waves from the transmitter to the observer also causes a form of transverse Doppler shift, for which Fig. 1(d) explains the operational conditions. If a transmitter T is placed in a vehicle, zero Doppler shift is observed, when the vehicle is located at a distance $\Delta S=(v/c)H$ from the zenith position at a time $\Delta T=H/c$ after passing the zenith.

DOPPLER SHIFT IN NONHOMOGENEOUS MEDIA

We have to use electromagnetic field theory in deriving rigorously the Doppler shift. The derivation is based on the following assumptions. A radiating system, given by a distribution of a current density $\overline{J}(\bar{r}')$, is located near the origin of a coordinate system. The consequence is an electromagnetic field which depends on the medium surrounding the radiator. In this field, an observer (receiver) is moving with a vector velocity $\bar{v}$. Its instantaneous position is given by $\bar{r}$. As shown previously, the observed angular frequency ω' is

$$\omega' = \omega + \frac{d\phi(\bar{r})}{dt} = \omega + \Delta\omega, \tag{1}$$

where ω is the frequency of the transmitter $[\exp(j\omega t)]$, ϕ is the phase angle of the observed field strength, and $\Delta\omega$ is the Doppler shift due to the motion of the receiver. We find

$$\Delta\omega = \frac{d\phi(\bar{r})}{d\bar{r}} \cdot \frac{d\bar{r}}{dt} = \nabla\phi\cdot\bar{v}. \tag{2}$$

This result is in agreement with that of rigorous treatments where the observer is placed in a second coordinate system moving with the vector velocity $\bar{v}$. The Doppler frequency obtained by the use of the Lorentz transformation (frequency in the moving frame) is

$$\omega' = (\omega - \nabla\phi\cdot\bar{v})/[1-(v/c)^2]^{1/2}.$$

The denominator in this equation takes into account a deviation due to relativity which is neglected above.

As a first step, we derive the observed field strength $\overline{E}(\bar{r})$, the phase of which gives the Doppler shift according to (2).

It may be noted that the results based on these assumptions also are applicable to the case of a mutually interchanged transmitter and receiver for nonrelativistic speeds and isotropic media due to the reciprocity theorem.

The properties of the medium are defined by a relative permittivity $\epsilon_r(\bar{r}')$ which may be complex, where

$$\epsilon_r(\bar{r}') = 1 + \Delta\epsilon_r(\bar{r}'). \tag{3}$$

Let us start with an evaluation of Maxwell's equations for the complex amplitude of the field magnitudes *E*, *D*, *H*, and *B* according to the following equations using customary procedures:

$$\nabla\times\overline{E} = -j\omega\mu_0\overline{H}, \tag{4}$$

$$\nabla\times\overline{H} = j\omega\epsilon_0\epsilon_r(\bar{r}')\overline{E} + \overline{J}, \tag{5}$$

$$\nabla\cdot\overline{B} = 0, \qquad \nabla\cdot\overline{H} = 0, \qquad \overline{B} = \mu_0\overline{H}, \tag{6}$$

$$\nabla\cdot\overline{D} = 0, \qquad \nabla\cdot[\epsilon_0\epsilon_r(\bar{r}')\overline{E}] = 0, \tag{7}$$

$$\epsilon_0\epsilon_r(\bar{r}')\nabla\cdot\overline{E} + [\nabla\epsilon_0\epsilon_r(\bar{r}')]\cdot\overline{E} = 0, \tag{8}$$

$$\nabla\times\nabla\times\overline{E} = -j\omega\mu_0\nabla\times\overline{H} = \omega^2\epsilon_0\mu_0\epsilon_r(\bar{r}')\overline{E} - j\omega\mu_0\overline{J}; \tag{9}$$

$$\nabla\times\nabla\times\overline{H} = j\omega\epsilon_0\epsilon_r(\bar{r}')\nabla\times\overline{E} + j\omega\epsilon_0[\nabla\epsilon_r(\bar{r}')\times\overline{E}] + \nabla\times\overline{J}. \tag{10}$$

Introducing the identity

$$\nabla\times\nabla\times\overline{E} = \nabla(\nabla\cdot\overline{E}) - \nabla^2\overline{E}, \tag{11}$$

we obtain

$$\nabla\times\nabla\times\overline{E} = -\nabla\left(\frac{\nabla\epsilon_r(\bar{r}')\cdot\overline{E}}{\epsilon_r(\bar{r}')}\right) - \nabla^2\overline{E}, \tag{12}$$

$$\nabla\times\nabla\times\overline{H} = -\nabla^2\overline{H}, \tag{13}$$

and finally

$$\nabla^2\overline{E} + \beta_0{}^2\overline{E} = j\omega\mu_0\overline{J} - \nabla\left(\frac{\nabla\epsilon_r(\bar{r}')\cdot\overline{E}}{\epsilon_r(\bar{r}')}\right) - \beta_0{}^2\Delta\epsilon_r(\bar{r}')\overline{E}, \tag{14}$$

$$\nabla^2\overline{H} + \beta_0{}^2\overline{H} = -j\omega\epsilon_0[\nabla\epsilon_r(\bar{r}')\times\overline{E}] - \beta_0{}^2\Delta\epsilon_r(\bar{r}')\overline{H}' - \nabla\times\overline{J}, \tag{15}$$

where $\beta_0{}^2=\omega^2\epsilon_0\mu_0$, with ϵ_0 and μ_0 for the permittivity and permeability of free space.

The wave equations (14) and (15) for the field distributions can be solved by diadic Green's functions. We solve for $\overline{E}$ at the position $\bar{r}$ of an observer and obtain an integral equation

$$\overline{E}(\bar{r}) = \int_V \overline{\Gamma}(\bar{r},\bar{r}')\cdot\left\{\overline{J} + j\,\frac{\omega\epsilon_0}{\beta_0{}^2}\,\nabla\left[\frac{\nabla\epsilon_r(\bar{r}')\cdot\overline{E}}{\epsilon_r(\bar{r}')}\right] + j\omega\epsilon_0[\Delta\epsilon_r(\bar{r}')\overline{E}]\right\}dV, \tag{16}$$

where

$$\overline{\Gamma}(\bar{r},\bar{r}') = \frac{1}{4\pi j\omega\epsilon_0}(\beta_0{}^2\mathrm{E} + \nabla\nabla)\,\frac{\exp(-j\beta_0|\bar{r}-\bar{r}'|)}{|\bar{r}-\bar{r}'|)} \tag{17}$$

is the diadic Green's function for the three-dimensional space.[3]

The final solution for $\overline{E}$ can be written as a series

$$\overline{E}(\bar{r}) = \overline{E}_0(\bar{r}) + \sum_{n=1}^{\infty}\Delta\overline{E}_n(\bar{r}). \tag{18}$$

The first term $[\overline{E}_0(\bar{r})]$ is the field distribution for empty space. It is

$$\overline{E}_0(\bar{r}) = \int_V \overline{\Gamma}(\bar{r},\bar{r}')\cdot\overline{J}(\bar{r}')dV. \tag{19}$$

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[3] H. Levin and J. Schwinger, "On the theory of electromagnetic wave diffraction by an aperture in an infinite plane conducting screen," in "Theory of Electromagnetic Waves, a Symposium," Interscience Publishers, Inc., New York, N. Y., pp. 1–37; 1951.

Introduction of (18) in (16) yields a recursion formula for the sum terms

$$\Delta\overline{E}_n(\bar{r}) = \int_V \overline{\Gamma}(\bar{r}, \bar{r}')\cdot \left\{ j\frac{\omega\epsilon_0}{\beta_0^2}\nabla\left[\frac{\nabla\epsilon_r(\bar{r}')\cdot E_{n-1}}{\epsilon_r(\bar{r}')}\right] + j\omega\epsilon_0[\Delta\epsilon_r(\bar{r}')\Delta\overline{E}_{n-1}]\right\} dV \quad (20)$$

where, for $n=1$, $\Delta\overline{E}_{n-1}=\overline{E}_0$.

Thus, the solution consists of a contribution by the incident waves $\overline{E}_0$ which would be obtained as a solution for $\Delta\epsilon_r=0$ and the sum of contributions due to waves scattered by the medium.

If $\Delta\epsilon_r \ll 1$, we can neglect terms of higher order and obtain

$$\overline{E}(\bar{r}) \approx \overline{E}_0 + \Delta\overline{E}_1. \quad (21)$$

Knowing $\overline{E}(\bar{r})$, we must proceed to evaluate it from the viewpoint of the application in (2). Let us write

$$\overline{E}(\bar{r}, t) = \overline{E}(\bar{r}) \exp(j\omega t) = \overline{A}(\bar{r}) \exp[j\phi(\bar{r})] \exp(j\omega t). \quad (22)$$

A known identity yields

$$\log_n A(\bar{r}) = \mathrm{Re}\left[\int_0^s \frac{dE(\bar{r})}{ds'} \Big/ E(\bar{r}) ds'\right] \quad (23)$$

and

$$\phi(\bar{r}) = \mathrm{Im}\left[\int_0^s \frac{dE(\bar{r})}{ds'} \Big/ E(\bar{r}) ds'\right]. \quad (24)$$

Introduction of (24) in (2) gives

$$\Delta\omega = \nabla\left\{\int_0^s \mathrm{Im}\left[\frac{dE(\bar{r})}{ds'} \Big/ E(\bar{r})\right] ds'\right\} \cdot \bar{v}. \quad (25)$$

We give s the direction of $\bar{v}$ and obtain

$$\Delta\omega \approx \mathrm{Im}\left[\frac{dE(\bar{r})}{d(\bar{v}t)} \Big/ E(\bar{r})\right] v. \quad (26)$$

Finally, if $\Delta\epsilon_r \ll 1$, it is as an approximation

$$\Delta\omega \approx \mathrm{Im}\left[\frac{d(E_0 + \Delta E_1)}{d(\bar{v}t)} \Big/ E_0\right] v. \quad (27)$$

The derivation is interesting, in so far as it shows that field theory permits straightforward calculation of the Doppler shift under complicated conditions of wave propagation. From this viewpoint, the above procedures may also be valuable as tools for the calculation of other problems.

Let us now use the derived equations (18), (20), and (27) to calculate some typical cases of the effect of the wave propagation on the Doppler shift.

EXAMPLE I—DEVIATION OF THE LONGITUDINAL DOPPLER SHIFT IN A VERTICALLY STRATIFIED MEDIUM

We consider as a first example the deviation of the longitudinal Doppler shift in a vertically stratified medium as it is indicated in Fig. 1(a) and described in the Introduction. Such deviations may occur and introduce errors when the Doppler shift is utilized for navigation in launching a space vehicle or in a terminal guidance system. If a profile of the properties of the medium is known, one can take these deviations into account and reduce the errors.

The following simplifying assumptions are made. We place a rectangular coordinate system with the X and Y axes on the ground as indicated in Fig. 2. Plane waves

$$E_x = \exp[j(\omega t - \beta_0 z)], \quad (28)$$

$$H_y = \frac{1}{Z_0}\exp[j(\omega t - \beta_0 z)], \quad (29)$$

$$\beta_0 = \omega\sqrt{\epsilon_0\mu_0}, \qquad Z_0 = \sqrt{\mu_0/\epsilon_0} \quad (30)$$

propagate in the positive Z direction. The medium is vertically stratified and defined by an eventually complex $\Delta\epsilon_r = f(z)$. The electric field strength is observed in the vehicle which travels with a velocity v_z in the vertical direction.

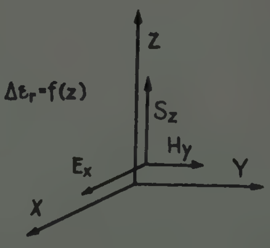

Fig. 2—Assumptions for Example I.

Under these conditions the incident field strength for $\Delta\epsilon_r = 0$ is

$$\overline{E}_0(z, t) = \bar{i}_x \exp[j(\omega t - \beta_0 z)]. \quad (31)$$

The properties of the medium are characterized by

$$\Delta\epsilon_r = f(z), \qquad \Delta\epsilon_r \ll 1, \qquad \nabla\epsilon_r = \bar{i}_z \frac{d\Delta\epsilon_r}{dz}\cdot \quad (32)$$

For ionized gas (ionosphere), we find[4]

$$\Delta\epsilon_r = -(\omega_c/\omega)^2 \quad (33)$$

where

$$\omega_c = 56\sqrt{N} \quad (34)$$

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[4] F. J. Tischer, "Wave Propagation Through Ionized Gas in Space Communications," Institute of the Aeronautical Sciences, New York, N. Y., IAS Rept. No. 59–34; January, 1959.

is the critical (plasma) frequency which depends, according to (34), on the electron density N (electrons per meter^3).

Considering the contributions to the field strength due to the medium (20), we find that the first term vanishes since $\nabla\epsilon_r$ and $\overline{E}$ are perpendicular to each other. We obtain

$$\Delta\overline{E}_1 = +j\omega\epsilon_0 \int_V \overline{\Gamma}(\bar{r}, \bar{r}')\cdot[\Delta\epsilon_r(z')\overline{E}_0]dV, \qquad (35)$$

$$\Delta\overline{E}_1 = \frac{1}{4\pi}(\beta_0{}^2\mathrm{E} + \nabla\nabla)\int_v \bar{i}_x\Delta\epsilon_r(z')$$

$$\cdot \exp\left[-j\beta_0(z' + R)\right]\frac{1}{R}\,rdrdz', \qquad (36)$$

where $r=(x^2+y^2)^{1/2}$, $R=[(z-z')^2+r^2]^{1/2}$, and $rdr=RdR$. Integration in the horizontal plane yields

$$\Delta\overline{E}_1 = \bar{i}_x\left[-\frac{1}{2}\beta_0 \exp(-j\beta_0 z)\int_0^z \Delta\epsilon_r(z')dz'\right]. \qquad (37)$$

We introduce (37) in (27) and obtain finally for the Doppler frequency

$$\Delta f \approx -\frac{\beta_0}{2\pi}[1 + \tfrac{1}{2}\Delta\epsilon_r(z)]v_z = -[1 + \tfrac{1}{2}\Delta\epsilon_r(z)]\frac{v_z}{\lambda_0}, \qquad (38)$$

where λ_0 is the free-space wavelength.

The relative deviation from the regular, free-space Doppler shift is then

$$S(z) = \tfrac{1}{2}\Delta\epsilon_r(z) = -\tfrac{1}{2}[\omega_c(z)/\omega]^2. \qquad (39)$$

Fig. 3 shows a diagram of the deviation vs frequency with the electron density N as parameter.

It may be noted that this result could also have been obtained directly from the known Doppler equation. It may serve to check the correctness of the derived equations.

EXAMPLE II—TRANSVERSE DOPPLER SHIFT IN A NONHOMOGENEOUS MEDIUM

Let us consider as a further example the transverse Doppler shift under the conditions shown in Fig. 1(b), where the direction of wave propagation is perpendicular to the direction of the velocity vector. This case is interesting since no Doppler shift would occur in free space. Such a Doppler shift must be taken into account in evaluating Doppler measurements on satellites for the determination of electron densities and in Doppler navigation.

As before, we assume plane waves according to (28)–(30) traveling from the ground plane (X, Y plane) in the positive Z direction. The medium is stratified in the Y direction and the velocity vector $\bar{v}$ of the observer has a Y component only:

$$\Delta\epsilon_r = f(y'), \qquad \bar{v} = \bar{i}_y v_y. \qquad (40)$$

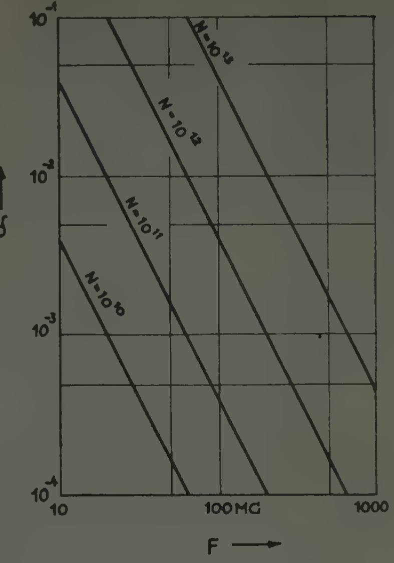

Fig. 3—Deviation of the Doppler shift in ionized gas vs frequency (N=electron density in particles per meter^3).

The incident field strength (for $\Delta\epsilon_r=0$) is

$$\overline{E}_0(z) = \bar{i}_x \exp(-j\beta_0 z). \qquad (41)$$

Considering the contribution to the electric field strength due to the medium (20), we find that the first term vanishes. We obtain

$$\Delta\overline{E}_1(z) = j\omega_0\epsilon_0\int_V \overline{\Gamma}(\bar{r}, \bar{r}')\cdot[\Delta\epsilon_r(y')\overline{E}_0(z)]dV, \qquad (42)$$

$$\overline{E}(z) = \overline{E}_0(z) + \frac{1}{4\pi}(\beta_0{}^2\mathrm{E} + \nabla\nabla)$$

$$\cdot\int_V \Delta\epsilon_r(y')\overline{E}\,\frac{\exp\left[-j\beta_0|\bar{r} - \bar{r}'|\right]}{|\bar{r} - \bar{r}'|}\,dV. \qquad (43)$$

Differentiation of (42) with respect to z yields

$$\frac{d\Delta\overline{E}(z)}{dz}$$

$$= \frac{1}{4\pi}(\beta_0{}^2\epsilon + \nabla\nabla)\iint\frac{\Delta\epsilon_r(y)\overline{E}_0(z)\exp(-j\beta_0|R|)}{R}\,dxdy. \qquad (44)$$

We assume that the variation of $\Delta\epsilon_r$ in the Y direction is small and expresses $\Delta\epsilon_r$ in the neighborhood of y_0 by

$$\Delta\epsilon_r(y) \approx \Delta\epsilon_r(y_0) + \frac{d\Delta\epsilon_r(y_0)}{dy}\Delta y, \qquad (45)$$

where $\Delta y = R\sin\phi$, $R=(x^2+y^2)^{1/2}$, and $dxdy = rd\phi dr$. The variation of $\overline{E}$ in the Y direction is assumed to be negligible in the region where $\Delta\epsilon_r$ contributes to the field strength. Integrating (44), we obtain

$$\frac{d\Delta\overline{E}}{dz} \approx -j\frac{1}{2}\beta_0\Delta\epsilon_r(y_0)\overline{E}_0; \qquad (46)$$

introducing

$$\frac{d\overline{E}_0}{dz} = -j\beta_0\overline{E}_0$$

yields the approximation

$$\frac{d\overline{E}}{dz} = -j\beta_0\left[1 + \frac{1}{2}\Delta\epsilon_r(y_0)\right]\overline{E}, \tag{47}$$

and finally

$$\overline{E}(y_0) \approx \overline{E}_0(0)\exp\left\{-j\beta_0\left[1 + \frac{1}{2}\Delta\epsilon_r(y_0)\right]z\right\}. \tag{48}$$

The Doppler shift is

$$\Delta\omega \approx -\frac{1}{2}\beta_0\frac{d[\Delta\epsilon_r(y)]}{dy}zv_y. \tag{49}$$

If $d[\Delta\epsilon_r(y)]/dy$ also varies in the Z direction, we can write as an approximation

$$\Delta\omega \approx -\frac{1}{2}\beta_0 v_y\int_0^z \frac{d[\Delta\epsilon_r(y)z]}{dy}dz. \tag{50}$$

For obtaining an idea about the magnitude of this transverse Doppler shift, let us consider a practical example after transforming (49) into

$$\Delta F = -\frac{1}{2}\frac{H}{\lambda_0}\frac{F_{c1}^2 - F_{c2}^2}{F^2}\frac{v}{\Delta L}, \tag{51}$$

where H is the height of a satellite ($H=500$ km), F is the operational frequency ($F=20$ mc, $\lambda_0=15$ meters), F_{c1} and F_{c2} are the cutoff frequencies at two locations $\Delta L(=1000$ km) apart. F_{c1} and F_{c2} may have the values 4 and 6 mc. For the velocity, $v=8$ km/second is chosen. We obtain

$$|\Delta F| = 7 \text{ cps.}$$

It can be seen that the effect decreases linearly with increasing frequency. Phase and frequency modulation can be expected due to this transverse Doppler shift.

The transverse Doppler shift is related to distortions of the phase front which may become serious in tracking satellites.

CONCLUSION

By formulating the Doppler effect mathematically in a general manner, one can derive relations for the Doppler shift under space flight conditions taking into account propagation effects which result mainly from nonhomogeneities of the medium along the flight path and along the path of the wave propagation. Ionospheric and interplanetary electron distributions can be considered as such nonhomogeneous media.

Field theory and the general formulation of the Doppler effect permit calculation of the deviations from the regular Doppler shift due to these propagation effects.

Of particular interest is a transverse Doppler shift which occurs when the flight path is perpendicular to the direction of wave propagation. No Doppler shift would be observed under normal conditions. Nonhomogeneous propagation causes such an irregular Doppler shift. It may be noted that a similar shift occurs due to relativity.

In space navigation, tracking, and exploration these and other effects must be taken into account. Theoretical calculations of the type shown can replace experimental investigations which, in space science, are extremely costly.

Communication Efficiency Comparison of Several Communication Systems*

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Summary—**Of increasing importance in communication systems, particularly those required for data transmission from space craft, is the minimization of the total transmitted energy required to transmit a given amount of data. Modulation systems can be compared on the basis of a figure of merit defined as the ratio of energy required per bit transmitted in the presence of a given noise spectral density. Using this criterion, several families of communication systems are compared in this article, including AM, FM, FM/FM, and orthogonal matched filter communication systems. It is found that the orthogonal systems uniformly require less power to transmit at a given information rate than other systems and approach closely the theoretical limit established by Shannon's channel capacity theorem.**

INTRODUCTION

A CONVENIENT way of comparing communications systems is to compare their efficiency in terms of β, the received signal energy required per information bit transmitted in the presence of a given uniform Gaussian noise spectral density.

$$\beta = \frac{E_{\min}}{\epsilon^2} \tag{1}$$

where

$E_{\min}$ = minimum received energy required per bit,
ϵ^2 = noise spectral power density (positive frequency).

Equivalently, β may be expressed as

$$\beta = \frac{P_{\min}}{\epsilon^2 H} \tag{2}$$

where

$P_{\min}$ = minimum received power required,
H = information rate (bits per second).

A lower bound on β may be derived from Shannon's channel capacity theorem

$$H = B \log_2 \left(\frac{P_{\min}}{\epsilon^2 B} + 1 \right)$$

$$\frac{H}{B} = \log_2 \left(\beta_0 \frac{H}{B} + 1 \right)$$

$$\beta_0 = \alpha(2^{1/\alpha} - 1) \tag{3}$$

where

B = channel bandwidth,
β_0 = channel limit of β for arbitrarily small probability of error,
$\alpha = B/H$.

From (3) it follows that a lower bound on β_0 is $\log_e 2 = 0.693$ which is attained as $\alpha \rightarrow \infty$, *i.e.*, for systems of unlimited bandwidth (and proper data encoding).

Eq. (2) may be rewritten in a form which facilitates β computations for various systems.

$$\beta = \frac{P_{\min}}{\epsilon^2 H} = \frac{P_{\min}}{\epsilon^2 B} \frac{B}{H} = \frac{S_i}{N_i} \alpha \tag{4}$$

where

S_i = input signal power,
N_i = input noise power.

Thus, for any given system, β may be computed from the input signal-to-noise ratio (for a desired error rate) and the ratio of bandwidth to information rate.

PULSE CODE MODULATION

As a detailed example of the method of computation of β efficiency, consider a PCM system where successive 0's and 1's are transmitted by means of phase modulation (that is, with equal power required for 0's and 1's). A maximum likelihood detector, where 0's and 1's are transmitted with equal probability, is one which divides the possible received signals into two sets such that the probability of recognizing the transmitted signal is maximized. For a detector whose output is $+A$ when a "0" is sent under noise-free conditions, $-A$ for "1", and is an odd function of signal plus noise between these limits, the maximum likelihood threshold is 0. Assuming band-limited Gaussian noise to be present in the output, the probability of error P_e is a symmetrical function of the transmitted signal, *i.e.*,

$$P(0|1) = P(1|0) = P_e = \Phi(-z) = \frac{1}{\sqrt{2\pi}} \int_{-\infty}^{-z} e^{-x^2/2} dx$$

where

$P(0|1)$ = probability that a 0 is received when a 1 is transmitted,
P_e = probability of error per bit,
z = output signal-to-noise voltage ratio (rms).

The receiver input signal-to-noise power ratio required to produce an output signal-to-noise power ratio of z^2

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depends on the exact modulation and detection system employed. For any modulation technique, a suppressed carrier coherent detection system requires the least input power. For such a system where the input RF signal spectrum has twice the bandwidth of the modulating signal, the required input signal-to-noise ratio is $z^2/2$, since coherent detection causes the information sidebands to add coherently.

From the sampling theorem, the information rate is, under the preceding assumptions,

$$H = 2(B/2) = B,$$

since the modulation bandwidth equals half the information bandwidth assuming that the error probability is small. Therefore,

$$\alpha = \frac{B}{H} = 1$$

$$\beta = \frac{S_i}{N_i}\alpha = \frac{z^2}{2}. \tag{5}$$

For example, if $P_e = 10^{-6}$, $z = 4.753$, $\beta = 11.30$. It should be noted that this resulting β is independent of the transmission bandwidth as long as suppressed carrier coherent detection systems are assumed. For example, if single sideband modulation is employed, α decreases to $\frac{1}{2}$, but the input signal-to-noise ratio required increases proportionately so that β remains the same.

However, β is significantly larger for noncoherent systems. For example, consider a noncoherent, PCM-AM system. Assuming that a low-pass filter restricts the modulating waveform to a bandwidth of $B/2$ and that each individual bit forms a sin x/x waveform, the input signal-to-noise power ratio S_i/N_i is

$$\frac{S_i}{N_i} = c\,\frac{S_s}{N_i},$$

where S_s is the peak input power of a single sin x/x pulse and c is the per unit number of ON pulses sent. Assuming that ON and OFF pulses (0's and 1's) are equally likely to be transmitted,

$$\frac{S_i}{N_i} = \frac{1}{2}\,\frac{S_s}{N_i}.$$

A maximum likelihood envelope detector is one where the decision threshold between 0's and 1's is such that $P(0|1) = P(1|0)$, as before. For a given bit error probability P_e, this occurs where

$$P_e = \int_s^\infty xe^{-x^2/2}dx = \int_0^y xe^{-(x^2+a^2)/2}I_0(ax)dx, \tag{6}$$

where

y = decision threshold,

$$a = \sqrt{\frac{2S_s}{N}}.$$

The integrand of the first integral of (6) is the probability density function of the envelope of narrow-band Gaussian noise. The integrand of the second integral is the probability density function of the envelope of a sine wave plus narrow-band Gaussian noise, where $I_0(ax)$ is the modified Bessel function.

For $P_e = 10^{-6}$, (6) gives $y = 5.257$, $a = 7.068$. For $c = 0.5$,

$$\frac{S_i}{N_i} = ca^2 = \frac{1}{2}(7.068)^2 = 24.98.$$

The input bandwidth is twice the data bandwidth B_d. Therefore,

$$\alpha = \frac{2B_d}{2B_d} = 1,$$

$$\beta = \frac{S_i}{N_i}\alpha = 24.98.$$

Thus, noncoherent, noncarrier-suppressed PCM requires slightly more than 3 db transmitted power to communicate at a given information rate than does a coherent PCM system assuming all other system parameters are held constant. This result assumes that the threshold is ideally established. In practical PCM systems, β often exceeds 100.

AMPLITUDE MODULATION

For a suppressed carrier, coherent detection double sideband amplitude modulation system, the output signal-to-noise ratio in a square bandwidth B_d is equal to twice the input signal-to-noise ratio (in a bandwidth $2B_d$). In other words,

$$\frac{S_i}{N_i} = \frac{1}{2}\,\frac{S_0}{N_0}. \tag{7}$$

The ratio of RF bandwidth B to information rate H is

$$\alpha = \frac{B}{H} = \frac{2B_d}{2B_dH'} = \frac{1}{H'}, \tag{8}$$

where H' is the number of bits transmitted per single sample.

From (4) and (7) it follows that β is

$$\beta = \frac{1}{2H'}\left(\frac{S_0}{N_0}\right). \tag{9}$$

To compute β where suppressed carrier, coherent detection amplitude modulation is used to transmit analog data, use can be made of Table I where H' can be found by interpolation of the entries for various signal-to-noise ratios, assuming that the signal is uniformly distributed through the modulating interval. Alternatively, use can be made of (84) in Appendix I.

For the transmission of digital data, using amplitude modulation, it is necessary to quantize the modulating waveform into a number of levels. From (89) in Appendix I, β may be written as

$$\beta = \frac{L^2 - 1}{6H'}z^2 \tag{10}$$

TABLE I

EQUIVALENT SIGNAL-TO-NOISE RATIO ($10 \log_{10} S/N$) FROM (84) FOR INFORMATION RATE H' DEFINED FROM TABLE III FOR A CONTINUOUS CHANNEL WHERE THE TRANSMITTED SIGNAL IS UNIFORMLY DISTRIBUTED OVER AN INTERVAL AND THE NOISE IN THE CHANNEL IS NORMALLY DISTRIBUTED. P_e' IS THE ERROR RATE PER SAMPLE IF THE CHANNEL WERE USED TO TRANSMIT L-LEVEL DIGITAL DATA (88)

L	P_e'	$10 \log_{10} S/N$	H' Bits per Sample
4	0.0830	11.05	1.791
8	0.0558	18.58	2.921
16	0.0452	25.21	3.973
32	0.0414	31.46	4.992
64	0.0399	37.58	5.998
128	0.0393	43.64	6.999
256	0.0390	49.68	8.000
512	0.0389	55.71	9.000
1024	0.0388	61.74	10.000

TABLE II

PROBABILITY OF ERROR PER SAMPLE P_e', (90), AND REQUIRED SIGNAL-TO-NOISE RATIO ($10 \log_{10} S/N$), (89), FOR A CONTINUOUS CHANNEL TO TRANSMIT DIGITAL DATA AT A BIT ERROR RATE OF 10^{-6} ASSUMING A UNIFORM DISTRIBUTION OF SIGNAL LEVELS AND THE LINEAR ADDITION OF NORMALLY DISTRIBUTED NOISE IN THE CHANNEL

L	P_e'	$10 \log_{10} S/N$	H' Bits per Sample
2	10^{-6}	13.54	1.000
4	2×10^{-6}	20.42	2.000
8	3×10^{-6}	26.56	3.000
16	4×10^{-6}	32.55	4.000
32	5×10^{-6}	38.50	5.000
64	6×10^{-6}	44.46	6.000
128	7×10^{-6}	50.42	7.000
256	8×10^{-6}	56.39	8.000
512	9×10^{-6}	62.36	9.000
1024	10×10^{-6}	68.34	10.000

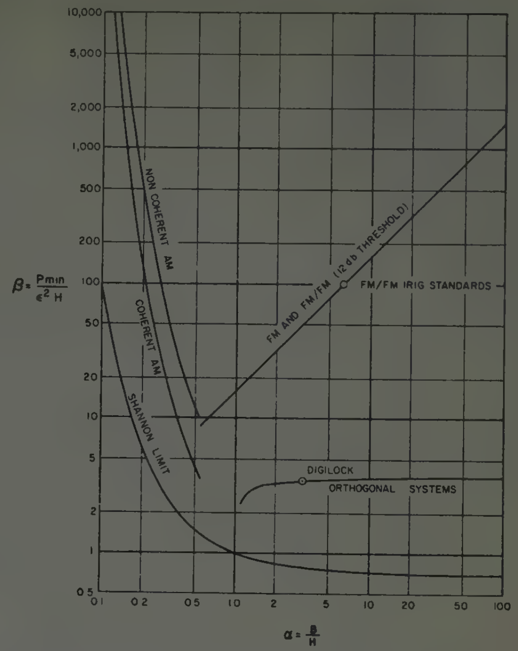

Fig. 1—B efficiency of various systems as a function of $\alpha = B/H$ for digital equivalent systems where rms quantizing noise equals rms error due to error rate ($\sigma_q = \sigma_e$).

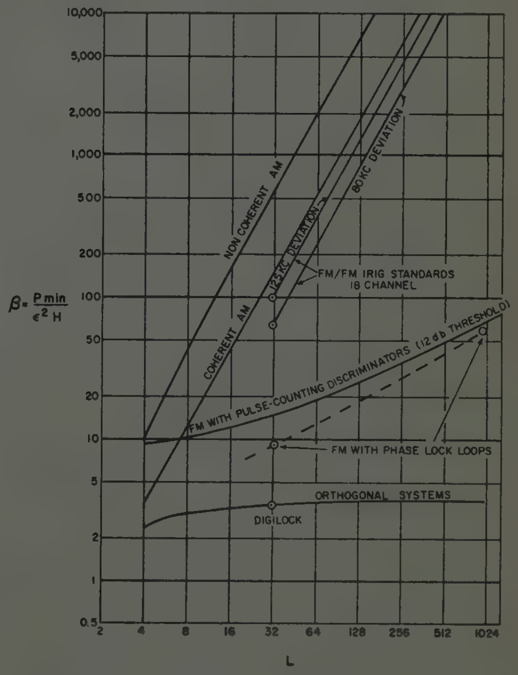

Fig. 2—B efficiency of various systems as a function of number of quantization levels L for digital equivalent systems where rms quantizing noise equals rms error due to error rate ($\sigma_q = \sigma_e$).

where, from (88)

$$P_e' = \frac{2(L-1)}{L}\,\Phi(-z). \tag{11}$$

In (10) and (11), L is the number of quantization levels and H' is defined by (85). The quantity z is defined by (11) for per sample error rate P_e'. $\Phi(-z)$ is the normal probability distribution function; *i.e.*,

$$\Phi(-z) = \frac{1}{\sqrt{2\pi}} \int_{-\infty}^{-z} e^{-x^2/2} dx. \tag{12}$$

For a bit error rate of 10^{-6} the signal-to-noise ratios shown in Table II may be used in conjunction with (9) to obtain β. For other error rates use may be made of (10)-(12).

Results of computations for amplitude modulation of both analog and digital data sources are shown in Figs. 1–4. Figs. 1 and 2 show systems transmitting analog data. Fig. 1 shows β as a function of α, while Fig. 2 shows β as a function of L, the number of levels which would be required to transmit the analog information with a digital system under the conditions for digital systems assumed in deriving Table III. Figs. 3 and 4 show systems as a function of α and the number of levels L if they are used for the transmission of digital data.

For noncoherent double-sideband amplitude modulation systems, the values of β are consistently higher than those for the coherent case for any given α. An approximate analysis of the noncoherent case assumes that a linear detector is used to determine the envelope of the incoming waveform. It is further assumed that

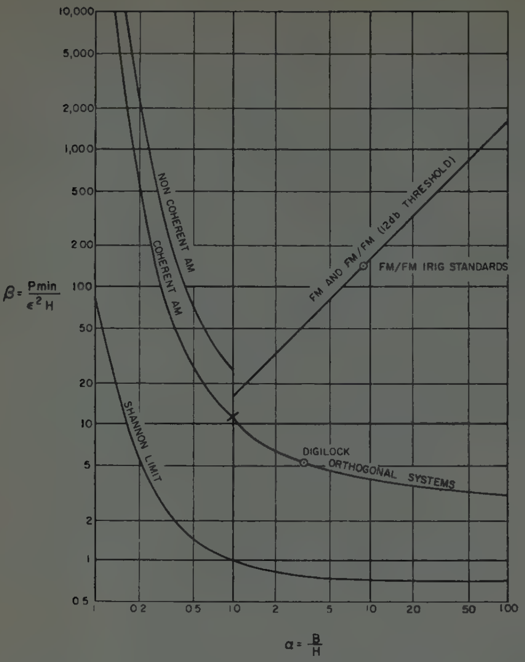

Fig. 3—B efficiency of various systems as a function of $\alpha = B/H$ for digital equivalent systems where the bit error rate $P_e = 10^{-6}$.

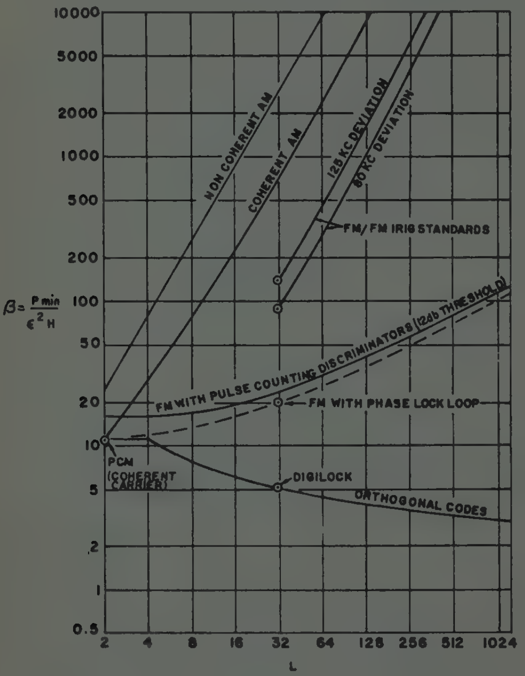

Fig. 4—B efficiency of various systems as a function of number of quantization levels L for digital equivalent systems where the bit error rate $P_e = 10^{-6}$.

TABLE III

EQUIVALENT SIGNAL-TO-NOISE RATIO ($10 \log_{10} S/N$) AND INFORMATION RATE H' (87) FOR QUANTIZED EQUAL-INCREMENT CHANNEL OF L-LEVELS WHERE RMS QUANTIZATION NOISE AND RMS ERROR DUE TO UNIFORM ERROR PROBABILITY P_e' ARE EQUAL

L	P'	$10 \log_{10} S/N$	H' Bits per Sample
4	0.025	8.75 decibels	1.791
8	0.00694	14.98 decibels	2.921
16	0.00184	21.06 decibels	3.973
32	4.73×10^{-4}	27.09 decibels	4.992
64	1.20×10^{-4}	33.11 decibels	5.998
127	3.03×10^{-5}	29.14 decibels	6.999
256	7.60×10^{-6}	45.16 decibels	8.000
512	1.90×10^{-6}	51.18 decibels	9.000
1024	4.76×10^{-7}	57.20 decibels	10.000

the input signal-to-noise ratio is sufficiently great so that the output noise is normally distributed about the modulating waveform as the mean value. As shown by (13)–(120) derived by Davenport and Root,[1] the output signal-to-noise ratio under high signal-to-noise ratio conditions is equal to twice the input signal-to-noise ratio for a linear detector. In other words, (7) can be assumed to hold approximately. However, the output signal contains a dc component which must be considered in arriving at the required input signal-to-noise ratio. In the coherent amplitude modulation case it was assumed that the signal was uniformly distributed over some interval, $-A/2$ to $A/2$, for instance. In the noncoherent case the interval normalized to the same units is 0 to A. The total output power for the coherent case is $A^2/12$, while it is $A^2/3$ for the noncoherent case. Therefore, the output signal power in the noncoherent case must be four times the output power in the coherent case for the same output noise. Hence β is

$$\beta = \frac{2}{H'} \frac{S_0}{N_0} \qquad \frac{S_i}{N_i} \gg 1. \tag{13}$$

Thus, the noncoherent amplitude modulation system used for the transmission of analog data is approximately 6 db less efficient than the coherent amplitude modulation system under high signal-to-noise ratio conditions.

For the transmission of digital data using noncoherent amplitude modulation, the output signal-to-nosie powers, assuming all levels are equally likely to be transmitted, are

$$S_0 = \frac{1}{L} \sum_{i=1}^{L} (\Delta A)^2 (i-1)^2 = \frac{(L-1)(2L-1)}{6} (\Delta A)^2, \tag{14}$$

$$N_0 = \sigma^2 = \frac{(\Delta A)^2}{4z^2}. \tag{15}$$

Assuming that (7) holds (*i.e.*, the input signal-to-noise ratio is large), (14) and (15) may be combined to give

$$\frac{S_i}{N_i} \simeq \frac{1}{2} \frac{S_0}{N_0} = \frac{(L-1)(2L-1)}{3} z^2. \tag{16}$$

α is given by (8); therefore, from (4),

$$\beta \simeq \frac{(L-1)(2L-1)}{3H'} z^2. \tag{17}$$

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[1] W. B. Davenport, Jr. and W. L. Root, "An Introduction to the Theory of Random Signals and Noise," McGraw-Hill Book Co., Inc., New York, N. Y.; 1958.

Thus, for the transmission of digital data the comparison of noncoherent and coherent amplitude modulation systems gives

$$\beta_{\text{noncoherent}} \simeq \frac{2(2L-1)}{L+1}\,\beta_{\text{coherent}} \qquad \frac{S_i}{N_i} \gg 1. \tag{18}$$

This result agrees closely with the more exact derivation of β for the particular case of noncoherent pulse code modulation given previously. In (18) for $L=2$, it is seen that β for the noncoherent system is twice that for the coherent system. The previous analysis showed that the ratio should be 2.21. It is also interesting to note from (18) that as the number of levels increase, the difference between noncoherent and coherent amplitude modulation systems approaches 6 db, the amount by which the noncoherent and coherent analog amplitude modulation systems differ. The β values for noncoherent amplitude modulation systems are plotted in Figs. 1–4.

FREQUENCY MODULATION

An analysis similar to the amplitude modulation case can be carried out for a frequency modulation system using conventional pulse counting discriminators. In the following discussion, formulas for transmission of analog data and of digital data are given together throughout the development since the derivations are essentially the same in both cases.

The output signal power for a signal uniformly distributed over the interval $-A/2$ to $A/2$ is $S_0=A^2/12$. In the frequency modulation case the output signal is proportional to the frequency deviation Δf of the carrier from the center frequency. Thus, for $A=\Delta f$,

$$S_0 = \frac{(\Delta f)^2}{3}. \tag{19}$$

For a uniformly distributed quantized signal the output power by (77) in Appendix I is

$$S_0 = \frac{(\Delta A)^2}{12}(L^2-1) = \frac{(\Delta f)^2}{3}\,\frac{L^2-1}{L^2} \tag{20}$$

where

$$\Delta A = \frac{2\Delta f}{L}.$$

The output noise power N_0 in the same units for large signal-to-noise ratios (*i.e.*, greater than the threshold value of approximately 16) in pulse type discriminators is

$$N_0 = \frac{\epsilon^2 B_d^3}{3S_i} \qquad \frac{S_i}{N_i} > 16, \tag{21}$$

where B_d is the data bandwidth, ϵ^2 is the noise spectral density, and S_i is the input signal power. Therefore, the output signal-to-noise ratios for continuous and quantized systems are the following:

continuous,

$$\frac{S_0}{N_0} = 2k\left(\frac{\Delta f}{B_d}\right)^3\left(\frac{S_i}{N_i}\right) \qquad \frac{S_i}{N_i} > 16; \tag{22}$$

quantized,

$$\frac{S_0}{N_0} = 2k\,\frac{L^2-1}{L^2}\left(\frac{\Delta f}{B_d}\right)^3\left(\frac{S_i}{N_i}\right) \qquad \frac{S_i}{N_i} > 16; \tag{23}$$

where

$$N_i = 2k\epsilon^2\Delta f,$$

$$k = \frac{B}{2\Delta f}.$$

Here, B is the RF bandwidth required for proper operation of the system so that k is the ratio of the actual RF bandwidth required to twice the carrier deviation. Assuming that the output signal-to-noise ratio for the quantized case is given by (89) in Appendix I, where z is defined by (88) or (90), the input signal-to-noise ratios required may be obtained by rewriting (22) and (23), making the suitable substitutions for the quantized case:

continuous,

$$\frac{S_i}{N_i} = \frac{1}{2k}\left(\frac{B_d}{\Delta f}\right)^3\frac{S_0}{N_0} \qquad \frac{S_i}{N_i} > 16; \tag{24}$$

quantized,

$$\frac{S_i}{N_i} = \frac{L^2z^2}{6k}\left(\frac{B_d}{\Delta f}\right)^3 \qquad \frac{S_i}{N_i} > 16; \tag{25}$$

where

$$P_e' = \frac{2(L-1)}{L}\,\Phi(-z). \tag{88}$$

The ratio of RF bandwidth to information rate α for both the continuous and quantized cases is

$$\alpha = \frac{2k\Delta f}{2B_dH'} = k\left(\frac{\Delta f}{B_d}\right)\frac{1}{H'}. \tag{26}$$

From (4), β for the two cases may be written as the following:

continuous,

$$\beta = \frac{1}{2H'}\left(\frac{B_d}{\Delta f}\right)^2\left(\frac{S_0}{N_0}\right) \qquad \frac{S_i}{N_i} > 16; \tag{27}$$

quantized,

$$\beta = \frac{L^2z^2}{6H'}\left(\frac{B_d}{\Delta f}\right)^2 \qquad \frac{S_i}{N_i} > 16. \tag{28}$$

It might appear from these equations that β can be made arbitrarily small by choosing the data bandwidth to be much smaller than the deviation; however, this can only be carried to the extent that the input signal-to-noise ratio remains above the frequency modulation threshold. From (24) and (25) it is seen that the ratio of data bandwidth to deviation is limited. Further, it might appear that β is independent of k; however, this is true only to the extent that the signal-to-noise ratio exceeds the threshold. Again from (24) and (25) the values of k are limited. If a frequency modulation system is operated at threshold, computation of β is comparatively simple. Assuming the threshold to exist at an input signal-to-noise ratio of 16, (4) shows that for both continuous and quantized cases,

$$\beta = 16\alpha, \quad \frac{S_i}{N_i} = 16. \tag{29}$$

For a given value of α, the minimum value of β exists at this threshold. Therefore, (29) indicates the best communication efficiency possible using conventional frequency modulation discriminators.

Values of β for analog and digital sources defined in the same way as for the amplitude modulation case, using Tables I and II, are shown in Figs. 1–4.

DOUBLE FREQUENCY MODULATION

For FM/FM systems using pulse counting discriminators, the preceding analysis for a single frequency modulation channel provides the basis of analysis for each FM/FM subcarrier. If the input signal-to-noise ratios can be computed for each subcarrier discriminator, the required input signal-to-noise ratio for the carrier discriminator can be computed by making the assumption that linear superposition of each subcarrier occurs. A frequency modulation discriminator, followed by a band-pass filter where the discriminator input is a carrier modulated by a sine wave of a frequency which falls within the filter bandwidth, provides an output signal-to-noise ratio of

$$\left(\frac{S_0}{N_0}\right)_j = 12\left(\frac{S_i}{N_i}\right)\frac{k\Delta f(\Delta f_j)^2}{(2k_j'\delta_{s_j})^3}\,\frac{1}{1+3\left(\dfrac{f_{s_j}}{k_j'\delta_{s_j}}\right)^2} \qquad \frac{S_i}{N_i} > 16, \tag{30}$$

where S_i/N_i is the input signal-to-noise ratio, Δf_j is the carrier deviation due to sine wave, ϵ^2 is the input noise spectral density, $2k_j'\delta_{s_j}$ is the band-pass filter bandwidth, and f_{s_j} is center frequency. The quantity k_j' is the ratio of bandwidth required in the band-pass filter to twice the subcarrier deviation for a particular subcarrier; $2k\Delta f$ is the total RF bandwidth at the input of the discriminator.

From (24) and (25) the required signal-to-noise ratios at the input of the jth subcarrier discriminator are the following:

continuous,

$$\left(\frac{S_i}{N_i}\right)_j' = \frac{1}{2k_j'}\left(\frac{B_{d_j}}{\delta_{s_j}}\right)^3\left(\frac{S_0}{N_0}\right)_j' \qquad \left(\frac{S_i}{N_i}\right)_j' > 16; \tag{31}$$

quantized,

$$\left(\frac{S_i}{N_i}\right)_j' = \frac{L_j^2 z_j^2}{6k_j'}\left(\frac{B_{d_j}}{\delta_{s_j}}\right)^3 \qquad \left(\frac{S_i}{N_i}\right)_j' > 16. \tag{32}$$

The primes in the preceding formulas indicate signal-to-noise ratios referred to the subcarriers. If the deviation of the main carrier by each subcarrier is reasonably small, superposition can be assumed to hold so that the output signal-to-noise ratio given by (30) is equal to the subcarrier input signal-to-noise ratios given by (31) and (32). Therefore, the required input signal-to-noise ratio for the main carrier is the following:

continuous,

$$\frac{S_i}{N_i} = \frac{k_j'^2}{3k}\,\frac{B_{d_j}^3}{\Delta f(\Delta f_j)^2}\left[1+3\left(\frac{f_{s_j}}{k_j'\delta_{s_j}}\right)^2\right]\left(\frac{S_0}{N_0}\right)_j' \qquad \frac{S_i}{N_i} > 16 \qquad \left(\frac{S_i}{N_i}\right)_j' > 16; \tag{33}$$

quantized,

$$\frac{S_i}{N_i} = \frac{L^2 z_j^2 k_j'^2}{9k}\,\frac{B_{d_j}^3}{\Delta f(\Delta f_j)^2}\left[1+3\left(\frac{f_{s_j}}{k_j'\delta_{s_j}}\right)^2\right] \qquad \frac{S_i}{N_i} > 16 \qquad \left(\frac{S_i}{N_i}\right)_j' > 16. \tag{34}$$

To compute the minimum value of β, it is again true that the system should be operated at threshold; however, thresholds now exist for not only the carrier discriminator but also for each subcarrier discriminator. An optimum FM/FM system is one where all thresholds are reached simultaneously. When this is the case, it is again true that $\beta = 16\alpha$. The quantity α for an FM/FM system with n subcarriers is

$$\alpha = \frac{k\Delta f}{\sum_{j=1}^{n} B_{d_j} H_j'}\,. \tag{35}$$

Although it is difficult to plot general curves for FM/FM systems, it is instructive to compute β for such a system assuming the IRIG standards apply. For these systems, $\delta_{s_j}/f_{s_j} = 0.075$ and $\delta_{s_j}/B_{d_j} = 5$, assuming a guard band of 30 per cent for each subcarrier $k_j' = 1.3$. Utilizing (31), it can be shown that at threshold for each subcarrier, the output signal-to-noise ratio is approximately 32 db. From Table I it is seen that this signal-to-noise ratio corresponds to approximately 5 bits per sample. Assuming that $\Delta f_j = c(f_{s_j})^{3/2}$ so that all subcarriers reach threshold at the same carrier signal-to-noise ratio, β may be computed using (35), assuming that the carrier discriminator threshold occurs at the same point as that for the subcarrier threshold. In conventional FM/FM systems, $k\Delta f$ is approximately 125 kc. Further, $\sum B_{d_j} = 4000$ cps for an 18-channel telemetry system. Using these figures, $\alpha = 6.25$, yielding a β of 100. Actually, with this carrier deviation the carrier discriminator reaches threshold at a higher signal level than that required for the subcarriers. Thus, the information rate is theoretically higher than 5 bits per sample in each subcarrier. However, in practice, due to cross-modulation products and system nonlinearities, the output signal-to-noise ratio at carrier threshold is not appreciably higher than that assumed. If the system is adjusted so that the carrier deviation is set for simultaneous thresholding of carrier and subcarrier discriminators, the theoretically best value of β is approximately 64.

FREQUENCY MODULATION WITH PHASE-LOCK-LOOP RECEIVERS

The analysis of this section is restricted to FM and FM/FM systems utilizing receivers with second-order phase-lock-loop filters. That is, referring to Fig. 5, the loop filter functions for a linearized phase-lock loop are[2]

$$F_1(s) = \frac{2\zeta + 2\zeta s}{Gs}, \tag{36}$$

$$F_f(s) = \frac{2\zeta^2}{s}. \tag{37}$$

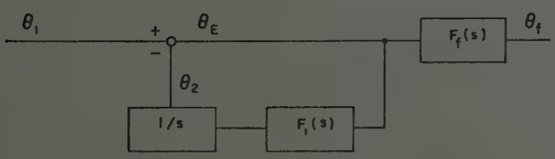

Fig. 5—Simplified block diagram of a phase-lock loop.

The resulting loop transfer functions are

$$Y_L(s) = \frac{\theta_2(s)}{\theta_1(s)} = \frac{2\zeta^2 + 2\zeta s}{2\zeta^2 + 2\zeta s + s^2}, \tag{38}$$

$$Y_E(s) = \frac{\theta_E(s)}{\theta_1(s)} = \frac{s^2}{2\zeta^2 + 2\zeta s + s^2}, \tag{39}$$

$$Y_f(s) = \frac{\theta_f(s)}{\theta_1(s)} = \frac{2\zeta^2 s}{2\zeta^2 + 2\zeta s + s^2}. \tag{40}$$

In the above equations

$$\zeta = \frac{4B_L}{3}, \tag{41}$$

where

$$B_L = \frac{1}{2\pi j}\int_0^{j\infty} |\, Y_L(s)\,|^2 ds. \tag{42}$$

For an input noise spectral density (normalized by signal power) Φ_n rad²/cps which is uniformly distributed in frequency (white noise), the mean square (steady-state) phase and frequency error for $\theta_1(t)=0$ are

$$\sigma_p^2 = \lim_{T\to\infty} \frac{1}{2T}\int_{-T}^{T} \theta_E^2(t)dt = \frac{1}{2\pi j}\int_{-j\infty}^{j\infty} \Phi_n |\, Y_L(s)\,|^2 ds$$
$$= \frac{3}{2}\,\Phi_n\zeta = 2\Phi_n B_L \quad \text{rad}^2 \tag{43}$$

$$\sigma_f^2 = \lim_{T\to\infty} \frac{1}{2T}\int_{-T}^{T} \theta_f^2(t)dt = \frac{1}{2\pi j}\int_{-j\infty}^{j\infty} \Phi_n |\, Y_f(s)\,|^2 ds$$
$$= \Phi_n\zeta^3 = \frac{64}{27}\,\Phi_n B_L^3 \quad \text{rad}^2/\text{sec}^2. \tag{44}$$

From (42), the noise bandwidth of the phase-lock loop is B_L. The 3-db bandwidth may be calculated from (39). It is found to be $\zeta\sqrt{2(2+\sqrt{5})} \simeq 0.6177 B_L$. However, the noise bandwidth and 3-db bandwidth of the frequency output of the phase-lock loop are not the same as for the phase output. For a constant deviation sine-wave modulated frequency source $\theta_1(t) = (\sin \omega_m t)/\omega_m$, the frequency response (power) is

$$\left| \frac{1}{s} \cdot Y_f(s) \right|^2.$$

Therefore, the noise bandwidth f_L of the frequency output of the phase-lock loop is

$$f_L = \frac{1}{2\pi j}\int_0^{j\infty} \left| \frac{4\zeta^4}{s^4 + 4\zeta^4} \right| ds = \frac{\zeta}{4} = \frac{B_L}{3}. \tag{45}$$

The 3-db bandwidth is

$$f_0 = \frac{\zeta\sqrt{2}}{2\pi} = \frac{2\sqrt{2}}{3\pi} B_L \simeq 0.3 B_L. \tag{46}$$

From the fundamental definition of β it follows that

$$\beta = \frac{1}{2\Phi_n H} = \frac{1}{4\Phi_n B_d H'}. \tag{47}$$

Using (43), this becomes

$$\beta = \frac{1}{2\sigma_p^2 H'}\left(\frac{B_L}{B_d}\right). \tag{48}$$

The restriction on the value of σ_p is dictated by the required proability of the phase-lock loop remaining in lock. Where the drop-out probability is P_p and an unmodulated carrier is present at the input,

$$\sigma_p^2 = \frac{\pi^2}{4z_p^2} \tag{49}$$

where

$$P_p = 2\Phi(-z_p) = \sqrt{\frac{2}{\pi}}\int_{-\infty}^{-z_p} e^{-x^2/2} dx.$$

The ratio B_L/B_d and the quantity z_p are restricted by the allowed transient error in the phase-lock loop. The phase of the incoming signal must change slowly enough to allow the loop to follow.

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[2] R. Jaffe and E. Rechtin, "Design and performance of phase-lock circuits capable of near-optimum performance over a wide range of input signal and noise levels," IRE TRANS. ON INFORMATION THEORY, vol. IT-1, pp. 66–76; March, 1955.

For an input signal which is a frequency ramp with slope $\Delta\dot{\omega}$ rad/sec^2,

$$\theta_1(t) = \tfrac{1}{2}(\Delta\dot{\omega})t^2,$$

$$\theta_1(s) = (\Delta\dot{\omega})/s^3.$$

The loop error signal $\theta_E(t)$ for this input is

$$\theta_E(t) = \frac{(\Delta\dot{\omega})}{2\zeta^2}\left[1 + \sqrt{2}e^{-\zeta t}\sin\left(\zeta t - \frac{3\pi}{4}\right)\right].$$

The maximum of this expression occurs at $\zeta t = \pi$. Labeling this maximum ξ, it is

$$\xi = \frac{(\Delta\dot{\omega})}{2\zeta^2}(1 + e^{-\pi}) \simeq \frac{\Delta\dot{\omega}}{2\zeta^2} \text{ rad.} \tag{50}$$

If the input signal is band-limited of bandwidth B_d with amplitude of frequency deviation uniformly distributed over an interval $-\Delta f$ to Δf, the signal waveform may be represented as a sum of sin x/x waveforms spaced $\frac{1}{2}B_d$ seconds apart. For two successive data points of amplitude $+\Delta f$ and $-\Delta f$, the peak slope of the waveform between the two data points is $16B_d\Delta f/\pi$. On the other hand, for a triangular waveform between the two data points, the slope is $4B_d\Delta f$. Therefore, from (50), the peak transient error ξ is bounded by

$$\frac{4\pi B_d\Delta f}{\zeta^2} < |\xi| < \frac{16B_d\Delta f}{\zeta^2} \tag{51}$$

for signals of amplitude limited to $\pm\Delta f$ and of bandwidth B_d. Assuming that the lower bound holds closely enough for purposes of analysis,

$$B_d\Delta f \simeq \frac{\zeta^2}{4\pi} = \frac{4}{9\pi} B_L{}^2\xi. \tag{52}$$

If B_d is chosen to be equal to the 3-db bandwidth of the frequency output of a phase-lock loop where ζ for the filter is $\zeta = 4B_L/3$, *i.e.*, $B_d = f_0$ in (46), it follows that

$$B_d\Delta f \simeq \frac{4}{9\pi}\left(\frac{3\pi B_d}{2\sqrt{2}}\right)^2 \xi = \frac{\pi}{2} B_d{}^2\xi$$

or

$$\frac{\Delta f}{B_d} \simeq \frac{\pi}{2}\xi\,. \tag{53}$$

Thus, for a peak allowable transient error $\xi = 0.5$ rad, the ratio of peak deviation to information bandwidth is restricted to values less than $\pi/4 = 0.785$. As will be shown, β for such a system is relatively high.

When phase modulation is present on the input signal, (49) must be modified to take the resulting transient error into account. If a transient error ξ' rad exists, the probability P_p' that the phase noise plus transient error is greater than $\pm\pi/2$ radians is

$$P_p' = \Phi\left(-z_p\left[1 - \frac{2}{\pi}\xi'\right]\right) + \Phi\left(-z_p\left[1 + \frac{2}{\pi}\xi'\right]\right).$$

For ξ' uniformly distributed over the interval $-\xi < \xi' < \xi$, the probability that the loop will fall out of lock is

$$P_p = \frac{1}{2\xi}\int_{-\xi}^{\xi}\left\{\Phi\left(-z_p\left[1 - \frac{2}{\pi}\xi'\right]\right) + \Phi\left(-z_p\left[1 + \frac{2}{\pi}\xi'\right]\right)\right\} d\xi'$$

or

$$P_p = \frac{1}{\xi}\int_{-\xi}^{\xi}\Phi\left(-z_p\left[1 + \frac{2}{\pi}\xi'\right]\right) d\xi'. \tag{54}$$

Although this expression is difficult to evaluate exactly, a useful lower bound exists. Since $\Phi(-z)$ and its first derivative are monotonic in magnitude decreasing functions of z (for $z > 0$), it follows that

$$\int_{z_1}^{z_2}\Phi(-x)dx > \left|\frac{\Phi^2(-z_1)}{2\Phi'(-z_1)}\right| = \frac{\Phi^2(-z_1)}{2\phi(z_1)}, \tag{55}$$

where

$$\phi(z) = \frac{1}{\sqrt{2\pi}} e^{-z^2/2},$$

under the conditions that

$$z_2 \geq z_1 + \frac{\Phi(-z)}{\phi(z)},$$

$$z_1 \geq 0.$$

Therefore,

$$P_p > \frac{\pi\left[\Phi\left(-z_p\left[1 - \frac{2}{\pi}\xi\right]\right)\right]^2}{4z_p\xi\phi\left(z_p\left[1 - \frac{2}{\pi}\xi\right]\right)}, \tag{56}$$

where

$$\frac{\pi}{2} \geq \xi \geq \frac{\pi}{4z_p}\,\frac{\Phi\left(-z_p\left[1 - \frac{2}{\pi}\xi\right]\right)}{\phi\left(z_p\left[1 - \frac{2}{\pi}\xi\right]\right)}.$$

Thus, P_p is the probability of losing lock if the transient error is uniformly distributed over an interval $-\xi$ to ξ and if the mean square phase noise is $\sigma_p{}^2 = \pi^2/4z_p{}^2$.

In order to compute β for phase-lock-loop systems, it is necessary to find an expression for H', the information transmitted per sample. This is most easily accomplished by utilizing the methods of Appendix I and the results tabulated in Tables I-III. For an input signal uniformly distributed in frequency in the interval $-\Delta f$ to Δf, the output signal-to-noise ratio, by (19) and (44), is

$$\frac{S_0}{N_0} = \frac{9\pi^2(\Delta f)^2}{16\Phi_n B_L{}^3} = \frac{9\pi^2}{4}\left(\frac{\Delta f}{B_L}\right)^3\left(\frac{S_i}{N_i}\right). \tag{57}$$

From (43), it follows that

$$\frac{S_0}{N_0} = \frac{9\pi^2}{8\sigma_p^2}\left(\frac{\Delta f}{B_L}\right)^2 \tag{58}$$

if B_L in (43) and (44) are equal. From (46) and (53) it follows that

$$\frac{S_0}{N_0} = \frac{\pi^2\xi^2}{4\sigma_p^2} = \xi^2 z_p^2. \tag{59}$$

Thus the product ξz_p is determined by the desired output signal-to-noise ratio and hence the desired information rate. Referring to Table I, it is seen that for $H'=5$, S_0/N_0 must be approximately 2000. Therefore, $\xi z_p = 44.7$. For a lock drop-out probability $P_p = 10^{-4}$, (56) may be solved to give $z_p = 30.64$ and $\xi = 1.459$ rad.

From (48),

$$\beta = \frac{1}{\sigma_p^2 H'}\left(\frac{B_L}{B_d}\right) = \frac{4z_p^2}{\pi^2 H'}\left(\frac{B_L}{B_d}\right)$$

$$= \frac{4(30.64)^2}{\pi^2(5.0)}\left(\frac{3\pi}{2\sqrt{2}}\right) = 254.$$

Such a high value of β is highly undesirable for space telemetry applications. It comes about because such a large proportion of the transmitter power must reside in the carrier or, equivalently, because the product ξz_p is so large. The question naturally arises as to methods of reducing β with phase-lock-loop systems.

One method is to modify the bandwidth of the frequency output filter $Y_f(s)$ so that the modulation index of the input signal is not restricted by (53). The set of equations which must be used to compute β in this case are summarized below:

$$\sigma_p^2 = 2\Phi_n B_L = \frac{\pi^2}{4z_p^2}, \tag{43), (49}$$

$$\sigma_f^2 = \frac{64}{27}\Phi_n B_f^3, \tag{60}$$

$$f_0 = B_d = \frac{2\sqrt{2}}{3\pi} B_f, \tag{61}$$

$$B_d \Delta f \simeq \frac{4}{9\pi} B_L^2 \xi, \tag{52}$$

$$\frac{S_0}{N_0} = \frac{9\pi^2(\Delta f)^2}{16\Phi_n B_f^3} = \frac{9\pi^2}{4}\left(\frac{\Delta f}{B_f}\right)^3\left(\frac{S_i}{N_i}\right), \tag{62}$$

or

$$\frac{S_0}{N_0} = \frac{9\pi^2}{8\sigma_p^2}\,\frac{B_L}{B_f}\left(\frac{\Delta f}{B_f}\right)^2, \tag{63}$$

$$\beta = \frac{1}{2\sigma_p^2 H'}\left(\frac{B_L}{B_d}\right). \tag{48}$$

Eqs. (60)–(63) are obtained from (44), (46), (57) and (58) by replacing B_L in the latter equations by B_f. B_f is defined by revising filter functions, (37) and (40), to

$$F_f(s) = \frac{2\zeta_f^2}{s}\,\frac{2\zeta^2 + 2\zeta s + s^2}{2\zeta_f^2 + 2\zeta_f s + s^2}, \tag{64}$$

$$Y_f(s) = \frac{2\zeta_f^2 s}{2\zeta_f^2 + 2\zeta_f s + s^2}\cdot \tag{65}$$

From the preceding equations, β may be found by simple algebraic substitutions. First using (43), (49), (52), (61), and (63) it follows that

$$\frac{S_0}{N_0} = \xi^2 z_p^2\left(\frac{B_L}{B_f}\right)^5, \tag{66}$$

or

$$\frac{B_L}{B_d} = \frac{3\pi}{2\sqrt{2}}\left(\frac{1}{z_p^2\xi^2}\,\frac{S_0}{N_0}\right)^{1/5}. \tag{67}$$

Therefore,

$$\beta = \frac{3z_p^2}{\pi\sqrt{2}H'}\left(\frac{1}{z_p^2\xi^2}\,\frac{S_0}{N_0}\right)^{1/5}, \tag{68}$$

where H' is defined from the output signal-to-noise ratio. For the analog case, (68) holds directly. For the transmission of discrete data, (68) is modified by (89) in Appendix I. In this case

$$\beta = \frac{3z_p^2}{\pi\sqrt{2}H'}\left(\frac{z_0^2}{z_p^2\xi^2}\,\frac{L^2-1}{3}\right)^{1/5}, \tag{69}$$

where z_0 is given by (88) or (90) depending on the encoding scheme.

For an output signal-to-noise ratio of 2000 and a drop-out probability, $P_p = 10^{-4}$, $\beta = 9.071$. This is a minimum value for (69) based on the foregoing assumptions including P_p evaluated from (56). Because (56) is a lower bound on P_p, the computed value of β is actually a lower bound. However, it is very close to the actual value. The other system constants which hold for this minimum value of β are $z_p = 4.311$, $\xi = 0.418$ rad, $\sigma_p = 0.364$ rad, $\Delta f/B_d = 8.57$, $B_L/B_f = 3.6147$, $\Delta f/B_L = 0.7115$, and $S_i/N_i = 5.29$ or 7.24 db.

For the transmission of digital data with $L = 32$, $P_e = 10^{-6}$, and $P_p = 2.0\times 10^{-7}$, $\beta = 20.55$. In this case $z_p = 5.9519$, $\xi = 0.370$ rad, $\sigma_p = 0.264$ rad, $\Delta f/B_d = 10.72$, $B_L/B_f = 4.30$, $\Delta f/B_L = 0.749$, $S_i/N_i = 9.59$ or 9.82 db. If P_e increases to 10^{-3} and P_p to 2×10^{-4} the resulting β changes to 8.31, corresponding to a signal-to-noise input ratio of 4.61 or 6.64 db.

An additional point has been calculated for the transmission of analog data where an output signal-to-noise ratio of 67 db is required. The resulting lower bound on β in this case is approximately 57.

Comparing these results with those computed for an FM system using conventional discriminators indicates that there is actually very little difference in the β efficiency of the two techniques. This fact may be seen by referring to Fig. 2.

It is difficult to compute the β efficiency of FM/FM systems utilizing phase-lock loops for both the carrier and subcarriers. In FM/FM systems employing conventional discriminators it is seen that β is approximately 6 db higher (using IRIG standards) than in a nonfrequency-multiplexed FM system. It is unlikely that double phase-lock-loop FM/FM systems will appreciably lower the value of β over conventional discriminator FM/FM systems. Two approaches are possible for carrier phase-lock-loop systems, namely, modulation-following and nonmodulation-following systems. For nonmodulation-following systems, such as those used in present-day narrow-band Microlock, 80 to 90 per cent of the power resides in the carrier due to the upper bound imposed on the carrier modulation index to obtain necessary linearity. Such systems, as discussed previously, result in very high values of β compared to modulation-following systems. For modulation-following phase-lock loops it is reasonable to assume that the required signal-to-noise ratio for proper system operation in the loop will be between 6 and 10 db. Making this assumption and assuming that the noise bandwidth of a phase-lock loop is roughly equal to the maximum carrier deviation, the resulting value of β for an output signal-to-noise ratio of 2000 can be no less than one-half that of the conventional discriminator FM/AM equivalent system; *i.e.*, β will not be less than approximately 30.

One alternative modulation scheme which shows promise for reducing β is a frequency multiplexed FM/AM system with suppressed or partially suppressed AM carrier. Double phase-lock loops could be used in this system without the requirement of large carrier to sideband power ratios.

ORTHOGONAL SYSTEMS

According to Shannon's channel capacity theorem, β should decrease as α increases for some communication systems. From Figs. 1 and 3 it is seen that this is true for amplitude modulation; however, α cannot be made greater than approximately 2 without some sophisticated encoding process. Also from the Figs. 1 and 3 it is seen that for FM systems β increases with increasing α. In this section a fairly simple set of orthogonal communication systems are shown to exist where β decreases with increasing α for arbitrarily large α.

In general, an orthogonal communication system is defined as one for which N messages can be transmitted, each of which consists of N degrees of freedom such that (discrete) matched filters may be designed which satisfy the following equalities:

$$\text{filter outputs} = \sum_{j=1}^{N} c_{ij}c_{kj} = \begin{cases} 0 & i \neq k \\ A & i = k, \end{cases} \tag{70}$$

where

c_{rs} = values of sth degree of freedom for the rth message or matched filter.

$r, s = 1, 2, \cdots, N.$

Examples of such systems are discrete pulse position modulation and discrete frequency modulation. Another example, discussed in more detail below, is a generalized PCM system. In all cases it is assumed that sufficiently accurate timing standards and filters are available at the transmitting and receiving sites to enable synchronism to be maintained. Assuming that code lock and carrier lock can be maintained, the β efficiency of an orthogonal system depends on the type of detector employed. Since for L equiprobable messages the system possesses L outputs, it is natural to select a maximum likelihood "greatest-of" type of detector.

From (70) it is seen that if bipolar signals are used, it is possible by means of a simple phase reversal to double effectively the number of levels for the same number of degrees of freedom if an output detector is constructed to detect both $+A$ and $-A$ outputs. In the following it will be assumed that there are N degrees of freedom with $L = 2N$ levels.

For square bandwidths, the bandwidth required to transmit N degrees of freedom f_n times per second is $B_n = f_nN/2$. Assuming P_e' and $L = 2N$ are given, information rate can be computed from (85) so that

$$\alpha = \frac{B}{H} = \frac{f_nN}{2f_nH'} = \frac{N}{2H'}. \tag{71}$$

For orthogonal systems the output signal-to-noise power ratio is N times the input signal-to-noise ratio. For a maximum-likelihood detector (see Appendix II) the input signal-to-noise ratio required to provide an error probability P_e' is

$$\frac{S_i}{N_i} = \frac{2z_N^2}{N} \tag{72}$$

where

$$\Phi(-z_N) \simeq \frac{P_e'}{2N-2} \quad P_e' \ll 1.$$

For this case,

$$\beta = \frac{z_N^2}{H'} \rightarrow \frac{z_N^2}{\log 2N} \qquad \text{for } P_e' \ll 1.$$

Since orthogonal systems are basically digital transmission systems, they must be treated somewhat differently from AM or FM systems; however, Table III, derived in Appendix I, still gives a good basis for comparison. For a desired output signal-to-noise ratio, a number of levels and associated error probability can be taken from Table III for the transmission of analog data. For the transmission of digital data, the error rate is given directly by (72). In the above formulas the number of levels L is equal to twice the number of degrees of freedom of the system; *i.e.*, $L=2N$.

One particular orthogonal system is especially attractive for space telemetry use. This is a generalized PCM system utilizing maximum redundancy Reed-Muller codes.[3]

APPENDIX I

COMPARISON OF DISCRETE AND CONTINUOUS SYSTEMS

The noise analysis of digital transmission systems generally depends on assigning some error probability per bit or per sample. In continuous systems it is convenient for comparison purposes to express rms errors in the same terms. A method of doing this is to derive an equivalent rms error for discrete systems made up of quantization error and error probability. For multilevel discrete systems with equal increments between levels, the variance of the quantization error is

$$\sigma_q^2 = \frac{(\Delta A)^2}{12} \tag{73}$$

where $\Delta A=$amplitude difference between levels. Further, if all levels of a particular system are equally likely to be transmitted and the error probability is uniformly distributed over all nontransmitted levels (*e.g.*, as in an orthogonal system), the variance of the error due to an error probability per sample P_e' is

$$\sigma_e^2 = (\Delta A)^2 \frac{P_e' L(L+1)}{6} \tag{74}$$

where $L=$number of quantization levels.[4]

Assuming that the system is designed so that $k\sigma_q^2=\sigma_e^2$ at the minimum permissible power level, it follows that

$$P_e' = \frac{k}{2L(L+1)} \cdot \tag{75}$$

The total variance of the error is $\sigma_t^2=\sigma_q^2+\sigma_e^2$ or

$$\sigma_t^2 = \frac{(\Delta A)^2}{12}(1+k). \tag{76}$$

In terms of ΔA, the signal power S, assuming all levels are equally probable, is

$$S = \frac{1}{L}\sum_{i=1}^{L}(\Delta A)^2\left(i - \frac{L+1}{2}\right)^2$$

$$= \frac{(\Delta A)^2}{12}(L^2-1). \tag{77}$$

Thus,

$$\sigma_t^2 = S\frac{1+k}{L^2-1} \cdot \tag{78}$$

In other words, the equivalent signal-to-noise power ratio of a digital system with L levels and error probability per sample P_e' given by (75) is

$$\frac{S}{N} = \frac{L^2-1}{1+k} \cdot \tag{79}$$

In the preceding equations, k is an arbitrary positive number which defines the ratio of error due to error probability and to number of quantization levels. For example, if $k=0$, (75), the error probability must be 0, and the signal-to-noise ratio is due solely to quantization error. If the quantization error and error due to error probabillty are equally effective in determining signal-to-noise ratio, $k=1$. Table III shows error probability and equivalent signal-to-noise ratio for various values of L from (75) and (79) for $k=1$.

Unfortunately, it does not follow that a comparison of information rate for continuous and quantized channels can be based on signal-to-noise ratio equivalents. Information rate for a continuous channel where the perturbing noise is independent of the transmitted signal is

$$H' = H'(y) - H'(n)$$

$$= -\int_{-\infty}^{\infty} p(y)\log_2 p(y)dy + \int_{-\infty}^{L} p(n)\log_2 p(n)dn \tag{80}$$

where

$H'(y)=$the entropy of the received signal including noise,

$H'(n)=$the entropy of the perturbing noise.

Assuming that the transmitted signal is uniformly distributed over an interval of $-A/2$ to $A/2$ and that the perturbing noise is normally distributed with zero mean and rms value σ, the probability distributions of the received signal and of the perturbing noise are

$$p(y) = \frac{1}{A}\left[\Phi\left(\frac{2y+A}{2\sigma}\right) - \Phi\left(\frac{2y-A}{2\sigma}\right)\right], \tag{81}$$

$$p(n) = \frac{1}{\sigma\sqrt{2\pi}}e^{-n^2/2\sigma^2}. \tag{82}$$

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[3] R. W. Sanders, "Digilock Telemetry System," presented at the National Symp. on Space Electronics and Telemetry, San Francisco, Calif.; September 28–30, 1959.

[4] Note that there are many ways of computing σ_e. It depends on the particular digital system being used and the signal and error distributions. The assumption used here of a uniform error distribution over all levels not transmitted applies to systems such as orthogonal systems. However, assuming that all levels are equally likely to be transmitted, the value of σ_e does not appear to be critically dependent on the type of digital transmission system used. For example, in a PCM system where the number of transmitted levels is a power of 2, $\sigma_e^2=(\Delta A)^2P_e(L^2-1)/3$ where P is the error rate per bit.

The entropy of the received signal corresponding to the first integral of (80) is

$$H'(y) = \log_2 A - \frac{\sigma}{A}\int_{-\infty}^{\infty} [\Phi(w+\gamma) - \Phi(w-\gamma)] \cdot \log_2 [\Phi(w+\gamma) - \Phi(w-\gamma)] dw \tag{83}$$

where

$$\gamma = \frac{A}{2\sigma},$$

$$\Phi(x) = \frac{1}{\sqrt{2\pi}}\int_{-\infty}^{x} e^{-x^2/2} dx.$$

The integral in this expression is difficult to evaluate analytically; however, its approximate value is calculable for large signal-to-noise ratios

$$H' \simeq \log_2 A - \log_2 \sigma\sqrt{2\pi e} + \frac{\sigma}{A}(2.6062)$$

$$\simeq \frac{1}{2}\log\frac{S}{N} + 0.7523\sqrt{\frac{N}{S}} - 0.2546 \qquad \frac{S}{N} > 5 \tag{84}$$

where

$$S = \frac{A^2}{12},$$

$$N = \sigma^2.$$

In order to compare continuous and discrete transmission systems on the basis of equal information rates, it is necessary to calculate the transmission rate per sample for a discrete channel. For such a system, which transmits L levels with an error probability P_e' per sample and where all levels are equally likely to be transmitted and all levels are equally likely to be chosen if an error occurs, the information rate per sample is[5]

$$H' = \log_2 L - P_e' \log_2 (L-1) + P_e' \log_2 P_e' + (1 - P_e') \log_2 (1 - P_e'). \tag{85}$$

For P_e' defined from (75), H' becomes

$$H' = \log_2 L - \frac{k}{2L(L+1)} [\log_2 (L-1)(2L^2 + 2L - k)]. + \log_2 (2L^2 + 2L - k) - \log_2 (2L^2 + 2L). \tag{86}$$

For $k=1$, this equation reduces to

$$H' = \log_2 L - \frac{1}{2L(L+1)} [\log_2 (L-1)(2L^2 + 2L - 1)] + \log_2 (2L^2 + 2L - 1) - \log_2 (2L^2 + 2L). \tag{87}$$

The value of H' given by (87) for a digital system, where the rms output errors due to quantization and due to transmission error rate are equal, is given in Table III. S/N in Table I has been computed from (84), assuming that the signal-to-noise ratio of a continuous channel is defined to provide transmission rates equivalent to the rates for the digital channels shown in Table III. For example, from the fourth entry in Table III it is seen that a signal-to-noise ratio of 27.09 db is required to transmit 4.992 bits per sample, the amount of information transmitted per sample in a digital system of 32 levels with an error rate P_e' of 4.73×10^{-4} per sample. It is seen, comparing the results of Table III and Table I, that a considerable difference exists in the output signal-to-noise ratios for continuous and discrete systems under the assumed conditions for equivalent information rates. Qualitatively, the reason for this is that a continuous channel makes comparatively small mistakes most of the time while the discrete channel makes comparatively large mistakes less frequently. This statement may be further elucidated by computing a probability of error for the continuous channel which can be compared with that for a discrete channel.

If a continuous channel is divided into L increments, each of magnitude ΔA, and the noise in the channel appears as the linear addition of Gaussian noise to the signal, the probability of error $P_e'(i)$ per sample for all levels except the first and last is

$$P_e'(i) = 2\Phi(-z) \qquad i = 2, 3, \cdots, L-1$$

where

$$z = \frac{\Delta A}{2\sigma},$$

$$\sigma^2 = \text{noise power},$$

$$\Phi(-z) = \frac{1}{\sqrt{2\pi}}\int_{-\infty}^{-z} e^{-x^2/2} dx.$$

For the first and last levels

$$P_e'(i) = \Phi(-z) \qquad i = 1, L.$$

Assuming all levels are equally likely to be transmitted, the average error rate per sample is

$$P_e' = \frac{2}{L}\Phi(-z) + \frac{L-2}{L}[2\Phi(-z)] = \frac{2(L-1)}{L}\Phi(-z). \tag{88}$$

Still assuming that all levels are equally likely to be transmitted and that the mean value of signal is zero, the average signal power is given by (77),

$$S = \frac{(\Delta A)^2}{12}(L^2 - 1).$$

Since $z^2 = (\Delta A)^2/4\sigma^2$, the output signal-to-noise ratio required for an error rate defined by (88) is

$$\frac{S}{N} = \frac{L^2 - 1}{3} z^2. \tag{89}$$

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[5] Based on the entropy definition of information. See C. E. Shannon and W. Weaver, "The Mathematical Theory of Communication," University of Illinois Press, Urbana, Ill.; 1949.

Using (89) to define z from the signal-to-noise ratios shown in Table I for the various discrete channels with L levels, P_e' for the continuous systems can be computed from (88). The results of such computations are given in Table I. From these computations the above qualitative statement may be interpreted more precisely. For example, again taking the fourth entry of Table I, it is seen that the continuous channel, if used in a discrete application, would commit an error in the discrete sense once each 24 samples. In the analog sense, each error, while significant, still has a high probability of being within a few standard deviations of the correct value. For the equivalent digital system an error other than quantization error is committed only once for each 2100 data samples; however, when an error is committed, there is no assurance that the resulting analog output is close to the transmitted value.

The preceding analysis provides a method for comparing analog and digital transmission of *analog* data. By interpolating between the entries of Table I, the comparison can be made by defining an information rate for the analog data based on a desired signal-to-noise ratio. This same information rate can be transmitted by a digital channel in a number of ways. By interpolating between the entries of Table III, a digital channel can be defined which will transmit at the same information rate as the analog system where it is assumed that the digital system error rate is fixed at a level which will cause the output analog error due to digital quantization to be equal to the analog error due to transmission error rate.

The analysis of analog and digital transmission systems may be completed by investigating more thoroughly the analog and digital transmission of *digital* data. An analog transmission channel may be used to transmit digital data by quantizing the analog channel into a given number of levels and adjusting the signal-to-noise ratio so that the probability of error between levels is a prescribed amount. Assuming that the levels are equidistant and that the noise in the output appears as a linear addition of Gaussian noise to a signal, the average probability of error per sample is given by (88). The required output signal-to-noise ratio is given by (89). Qualitatively, this comparison of analog and digital systems assumes that any error is equally bad and that information rate for either type of transmission is defined in terms of channel error probability.

In order to define error rate in terms of error per bit for digital transmission using quantized continuous channels, it is necessary to postulate the method of assigning binary numbers to each level. The smallest bit error rate occurs if a Gray code is used so that adjacent levels differ by only one binary position. In this case, assuming that the per sample error rate is small, the error per bit P_e is

$$P_e = P_e'/\log_2 L.$$

Therefore, bit error rate for an analog channel transmitting Gray coded digital data is

$$P_e = \frac{2(L-1)}{L \log_2 L}\,\Phi(-z). \tag{90}$$

Values of analog signal-to-noise ratios required for a bit error rate of 10^{-6} are shown for various numbers of levels in Table II.

Appendix II

Maximum Likelihood Detector Orthogonal System Error Rates

From the sampling theorem, there are $N=2BT$ degrees of freedom in a message of bandwidth B and of duration T. For an orthogonal system of $L=2N$ levels, the signaling rate for small error probability P_e' is

$$H = \frac{1}{T}\log L = \frac{4B}{L}\log L.$$

For a fixed bandwidth, the signaling rate for $L=2$ is equal to the rate for $L=4$. Therefore, a 4-level orthogonal system is equivalent to a 2-level PCM system. In a linear PCM system with two equiprobable states, a maximum likelihood detector is one which is defined by a suitably designed threshold. Assuming a "0" sent is represented by $+S$ volts and a "1" by $-S$ volts, the maximum likelihood threshold is 0 volts. Any detector output which is positive is taken to be a "0"; negative outputs are taken as "1's." In a 4-level orthogonal system there are two output busses. If signal components s_1, s_2 and noise components n_1, n_2 are received, the output from each bus γ_1 and γ_2 is

$$\gamma_1 = s_1 + s_2 + n_1 + n_2. \qquad s_1, s_2 = \pm S$$
$$\gamma_2 = s_1 - s_2 + n_1 - n_2.$$

The outputs from γ_1 under noise-free conditions are $2S$, 0, 0, $-2S$ for codes 00, 01, 10, 11 in. The corresponding γ_2 outputs are 0, $2S$, $-2S$, 0. The error performance of the system is independent of the code which is sent. Assuming 00 is transmitted, an error will be committed by a "greatest-of" detector; *i.e.*, a detector which chooses the largest buss output magnitude at the most probable transmitted signal, if and only if

$$2S + n_1 + n_2 < |n_1 - n_2|. \tag{91}$$

This inequality holds only if $n_1 < -S$ or if $n_2 < -S$ or both. Therefore, a 4-level orthogonal system with a "greatest-of" detector is equivalent to a PCM system with a maximum likelihood detector.

For a PCM signal disturbed by Gaussian noise, the probability of an error P_{e_1} being made is $\Phi(-Z)$ where z is the signal-to-noise ratio S/σ^2 and $\Phi(-z)$ is the normal distribution function. Thus, the probability of error P_{e_2} for a 4-level orthogonal system is

$$P_{e_2} = 1 - \Phi^2(z).$$

Therefore, from (91),

$$\Pr(2S + N_{21} \cdot < |N_{22}|) = 1 - \Phi^2(z), \qquad (92)$$

where

$$N_{21} = n_1 + n_2,$$
$$N_{22} = n_1 - n_2.$$

If n_1, n_2 are independent random variables of standard deviation σ, N_{21} and N_{22} are also independent random variables but with standard deviation $\sigma\sqrt{2}$. Letting $S^* = 2S$, $N_1^* = N_{21}$, and $N_2^* = N_{22}$, it follows from (92)

$$P_{e_2} \Pr(S^* + N_1^* < |N_2^*|) = 1 - \Phi^2\left(\frac{z^*}{\sqrt{2}}\right), \qquad (93)$$

where

$$z^{*^2} = \frac{4S^2}{2\sigma^2} = E\left(\frac{S^{*^2}}{N_1^{*^2}}\right) = 2z^2.$$

For an L-level orthogonal system $(L-2N)$, the error probability for a "greatest-of" detector is

$$P_{e_N} = 1 - \int_0^\infty \phi(x - z_N^*)\,[\Phi(x) - \Phi(-x)]^{N-1}dx \qquad (94)$$

where

$$z_N^* = z\sqrt{N}.$$

By expanding (94) and making use of (93) it follows that

$$P_{e_N} \leq 1 - (N-1)\Phi^2\left(\frac{z^*}{\sqrt{2}}\right) + (N - z)\Phi(z^*)$$

$$\simeq 2(N-1)\Phi\left(-\frac{z^*}{\sqrt{2}}\right) z^* \gg 1.$$

Therefore,

$$\alpha = \frac{B}{H} = \frac{N}{2\log 2N} = \frac{L}{4\log L} \qquad (L > 2),$$

$$\beta = z^2\alpha = \frac{z^{*^2}}{2\log 2N} = \frac{z_N^2}{\log 2N} = \frac{z_N^2}{\log L}, \qquad (95)$$

where

$$z_N = \frac{z^*}{\sqrt{2}},$$

$$\Phi(-z_N) = \frac{P_{e_N}}{(2N-1)} = \frac{P_{e_N}}{L-2} \qquad (L > 2).$$

A plot of (95) is shown in Fig. 6. By a detailed analysis of (94) it can be shown that β approaches $\log_e^2$ as n approaches infinity for an arbitrarily small P_{e_N}.

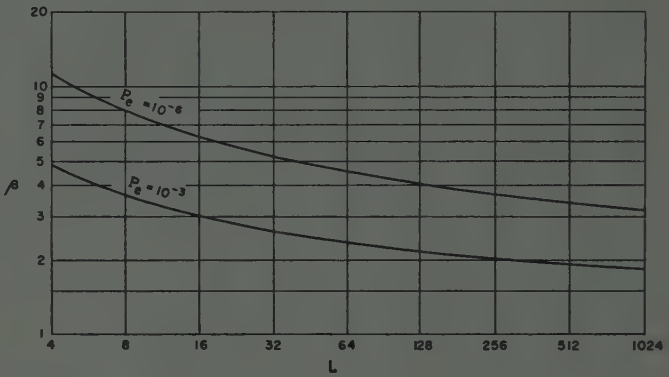

Fig. 6—Communication efficiency β of the orthogonal systems as a function of the number of levels L for bit error rate P_e of 10^{-3} and 10^{-6}.

Maximum Utilization of Narrow-Band Data Links for Interplanetary Communications*

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Summary—**Our ability to communicate with vehicles in space continues to be limited by the amount of received energy required to transmit a unit of information to earth. We have available, in practical hardware, radio-communication systems which provide us with theoretical maximum performance in terms of data bandwidth and distortion for a given transmission range and power output. The development and refinement of transmission, receiving, and power-supply techniques will extend the range of communication or increase the data bandwidth at a given range; but any significant extension of our capacity to communicate in space must come from advances in the state-of-the-art of telemetry data handling. New and advanced means for data processing and encoding prior to transmission should be investigated.**

The discussion which follows defines some technical areas which merit investigation, presents an outline for one approach to such an investigation and contemplates the evaluation of findings.

INTRODUCTION

THE PROBLEM facing those wishing to make physical or physiological experiments in unusual environments is the collection, transmission and interpretation of experimental data. The problem extends from the collection of data at primary sensors, through a sequence of data processing and transmission steps, and ends at the final interpretation of the experimental result, usually in a written report. Current practice in this process is well known and need not be reiterated here; however, we do not know where, in detail, we may improve on current practice. One area which seems to promise the next great advance in telemetry is the detection and coding for efficient transmission of basic data, at low-energy levels, which has been severely contaminated by noise.

An evaluation of the present state-of-the-art of sensors and data links is of interest in determining requirements or opportunities for improvements, based on detection and signal processing. We may postulate: if data links of any desired performance (bandwidth, range, linearity, distortion) were available, there would be no requirement for data detection prior to transmission and only distortionless amplification would be required. That is, if we could receive at the fixed ground station all of the original sensor output, we would be able to process the data at leisure, bringing to bear all the most advanced technology of data processing (digital computers). The foregoing statement does not remove the necessity to study methods for recovery of the desired data from relatively intense noise. A requirement for the development of advanced sensors and data links exists, but the characteristics of available data links will determine what portion of the data processing techniques to be developed should profitably be accomplished prior to data transmission by the telemetry radio link.

If such processing functions can be accomplished in reliable equipment with suitable environmental immunity, it appears possible to achieve additional gain in information from experiments which are currently limited by the capabilities of the radio link. Conversely, we may be able to perform, with a given vehicle which is weight- or power-limited, additional types of experiments which also result in a net increase in the amount of knowledge gained from each vehicle.

DATA HANDLING SEQUENCE

The information flow for a single experiment is illustrated in Fig. 1. In this figure, the progression of data from sensor to report is defined by the independent variable, x. The on-board (preliminary) and ground data reduction steps are separated by a data link, the only function of which is to transfer all of the information in existence at x_0 from one physical location to another.

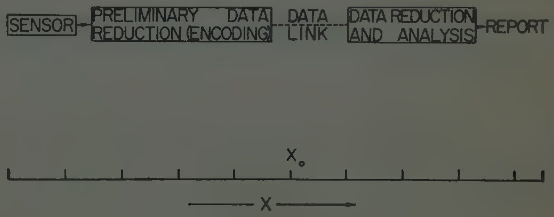
SENSOR
PRELIMINARY DATA REDUCTION (ENCODING)
DATA LINK
DATA REDUCTION AND ANALYSIS
REPORT
X₀
X

Fig. 1—Information flow.

As the signals, which contain the results of the experiment, progress down the line of information flow, they are modified progressively to enhance the desired data and to average out random errors and correct for identifiable systematic errors. Throughout this process, the information content, in the entropy sense, may be considered to be constant and equal to the information desired for the final report. Any redundant data carried along at any point in the process represent unnecessary data as far as the final report is concerned, and the transmission of such redundant data on the data link results in a demand for extra bandwidth over the minimum requirement for the fundamental information. We hope, in addition, that the redundancy content is at least equal to the noise and error content so that this redundancy can be used, as the data reduction process

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continues, to remove the noise and enhance the signal to the desired degree. Thus, for a potentially successful experiment, the bandwidth required at any given transfer point x_0 must be equal to or greater than the remaining signal redundancy plus information entropy.

ELEMENTS FOR PRELIMINARY INVESTIGATION

The characteristics of input transducers or sensors must be studied and tabulated. The nature of the output signal and the manner in which this signal represents the final desired data must be understood. We should search for potential sources of noise and systematic errors and obtain mathematical descriptions of these, where feasible. A detailed knowledge of these factors is essential before any discussion of optimum data reduction processes.

There may be certain physical limits to the detectability of the desired final data. One obvious limit is the quantum effect, which probably does not come into play in the majority of the experiments to be considered. Another limit is the result of random noise and is usually expressed in terms of noise bandwidth compared to signal bandwidth or information content. This concept is not limited to linear processes but can be extended to nonlinear processes as well, provided appropriate definitions are made of noise bandwidth and information content. These, and other, physical limits to detectability must be carefully stated and understood in order to avoid attempting the impossible. This is one of the most important "negative" results of information theory, one which has been of great importance in the practical design of communication systems.

We may define two classes of signal handling processes, one of which is allowable for inclusion in an optimum system. This first class is characterized by the fact that the ratio of noise power to remaining redundancy and information content is never greater at the output than at the input. In other words, the basic detectability of the desired data is unchanged. A perfect linear amplifier is obviously a member of this class, as are the usual correlation techniques. The net result of processing a signal through a sequence of such allowable operations is the recovery of the desired information with maximum signal-to-noise ratio as defined by the basic statement of detectability.

The second class of signal handling processes violates the requirement that the detectability of the signal be not reduced by the inclusion of the process. One such process is square-law or envelope detection of a signal at less than unity signal-to-noise ratio. Certain digital sampling processes may also be forbidden.

By investigating and tabulating the allowed and nonallowed signal handling processes, we hope to gain an insight into practical system designs for optimum detection. Of particular importance is the "negative" result than any optimum data-processing device must not include any process of the second class.

In the development of practical detection and encoding equipment and in the integration of such equipment into an optimum data system, an understanding of the basic characteristics of data links is a prerequisite. As stated previously, the present theoretical and practical limits of data links are the only justification for any consideration of additional data handling equipment to be included in the vehicle prior to data transmission. In fact, these current limitations are the determining factor in any decision relative to how much equipment is to be included in the vehicle.

The foregoing preliminary investigations would serve to summarize and codify present knowledge in a form applicable to the problem at hand. Some of the basic results of information theory, when restated in appropriate form for ready reference, would serve as a guide to the identification of those portions of the present art which already approach theoretical maximum performance. Concurrently, we should expect to identify areas wherein greatest improvement is feasible, because present art falls significantly short of predicted potential performance. For the direction of further studies, primary efforts should be concentrated where the greatest gain is available.

CONCEPTS OF OPTIMUM DATA REDUCTION

Reduced Redundancy

There is obviously no requirement for duplicate or redundant statements of the same datum by the time information appears in the final report, which is the end of the experiment—data reduction sequence. By lack of redundancy in the final statement of the datum, we mean that any further boiling down or condensation of the statement would change its meaning in an irretrievable manner. If such a nonredundant statement of a datum has been achieved by a series of allowable data reduction processes which do not degrade the theoretically attainable signal-to-noise ratio, then we must have realized the maximum possible improvement in the data. This principle can be used as a guideline to produce practical equipment for optimum data reduction, and the principle serves to identify the endpoint of the process.

We must recognize that a datum is still redundant if it contains any *a priori* information. Reducing the redundancy of a statement by removal of previously known facts may change the visible character of the statement, but not irretrievably, since we may always reconstruct that which we already knew before the experiment was performed. The maximum utilization of *a priori* information is the keystone to any correlation-detection process.

Although we may boil all the redundancy out of a datum, its value as information may not be very high if we could have predicted with high probability the result of the experiment. The information content is actually inversely proportional to the probability of occurrence of the observed datum. Surprises are worth the struggle, but mere reconfirmation of an expected result is of little value. Consequently, an aim of data reduction prior to transmission or prior to reporting should be to smooth out or equalize the surprise content of each datum communicated. We can most efficiently use the communication channel by transmitting continuously, at maximum rate, data which are uniformly high in information content. (It should be noted that the final written report is probably the narrowest information channel in the whole chain.)

In our zeal to wring all possible redundancy from a datum before reporting it, we may go too far. A completely nonredundant statement may well be too compressed for understanding by anyone other than the designers of the experiment. As an example, suppose we are to report electrocardiograms from deep space. The correlation detection might be in the form of a set of matched filters which produce output pulses, sorted to appropriate wires, with each characteristic beat of the heart. The transmitted data might be a series of numbers indicating the frequency of occurrence of pulses from each of the filters. Pulses representing normal heart operation would be merely reported as present, but abnormal or "surprise" pulses would be reported in detail. Such a report contains all the data in a nonredundant form and has been arranged to appropriately weight surprise data; however, the report may be unintelligible to an experienced doctor who would prefer to read the actual ECG waveforms and make his own interpretation from experience.

It is certainly feasible to regenerate "example" waveforms based on the nonredundant received data. The normal pulses would be written repeatedly at the reported rate, and the abnormal pulses interjected as instructed by the received data. Such an example waveform would not be guaranteed to be identical to the original gauge output, but it would contain exactly the same nonredundant data. The necessity or desirability of such regeneration of "example" waveforms is based on our own psychological desire to approximate the physiological impressions we might have received by performing the experiment directly, thereby gaining a better understanding of what occurred.

The maximum-likelihood concept of correlation detection might be described as follows. Transmitter and receiver are assumed to have a store containing all possible messages. On receipt of a signal contaminated by noise, the receiver compares (correlates) the received information with its store of possible messages and accepts that message which exhibits the highest correlation as the one that was most likely transmitted. The output datum is then the (nonredundant) one number identifying the maximum-likelihood message.

This concept is excellent, provided we can conceive in advance all possible outcomes of the proposed experiment. However, there is great danger that we may render ourselves incapable of recognizing surprise information that would have been of much greater value than any of the messages (experimental results) we might have predicted. This was exactly true in the radiation experiments in the early Explorer satellites. Only expected radiation levels were accounted for in the design of counters and data channels. Consequently, when the satellite encountered much higher radiation intensities, the counter mechanism was completely blocked and an incorrect indication of zero radiation was transmitted. The correct interpretation of these data took a lot of head scratching and several more satellites with more sophisticated and wider range counters to provide the right answer. The lesson to be learned from this example is that we must include in our correlation detectors, and in any other advanced data reduction devices, the ability to recognize that *none* of the *a priori* (expected) results is true. This would take the form of a confidence indication which would add information relative to the actual degree of correlation achieved by the reported maximum-likelihood message.

An added degree of sophistication might be achieved by automatic data handling equipment which recognizes experimental results and continuously adjusts the parameters of the detection process for best results in a changing situation. The storage of voltages, representing present frequency and rate-of-change of frequency, on capacitors in a third-order phase-locked loop, is a simple example of this degree of sophistication. The effect of certain types of automatic gain control or narrow band limiting on noise bandwidth is another example of such adaptive performance, utilizing *a posteriori* data.

It is apparent that any nonredundant datum must be transmitted without distortion or the meaning will be irretrievably lost. This fact has led to intensive effort in the development of error-correcting codes for the transmission of digital data. Such codes actually amount to adding redundancy for the purpose of transmission reliability. Spectrum expansion by pulse-code modulation or frequency modulation is also a means for adding redundancy to improve the reliability of the radio link.

Since the radio link may require redundancy for satisfactory operation, it appears feasible to provide some or all of the required redundancy in the data to be transmitted. In operation, this concept involves averaging of radio noise as well as transducer noise in the final data reduction process.

Statistics; Identification of Signal and Noise

In any filtering or data reduction problem, we are forced to state in advance those characteristics by which we will identify the signal components among the signal and noise information actually received. These *a priori* characteristics can usually be stated only in a statistical sense, and the most common analyses are based on stationary (nonchanging) statistics.

We often have a nice intuitive feeling that "signal" must be that portion of received information that best fits some smooth mathematical function and "noise" must be that part that seems to be random and rapidly changing. Indeed, this is the basis for the very earliest treatment of experimental errors by "least-squares" curve fitting techniques. The work of Norbert Wiener has extended this concept to include completely statistical representations of both signal and noise. His work has allowed the construction of least-square optimum filters for real-time data analysis. These Wiener filter techniques are well known and can be applied to any problem, the object of which is to recover the actual signal waveform in real time. Please note, however, that we can do somewhat better by "post flight" data reduction in which "past" as well as "future" data can be used and finite integrals are permitted. Similar results can be obtained in almost real time if time delay is permitted and appropriate storage mechanisms are included.

The proper application of such statistical filtering techniques for the recovery of actual signal waveforms should play an important part in any study of data processing techniques. However, truly significant gains in data analysis will be accomplished only if further reduction of redundancy is accomplished. This might take the form of the computation of values for the various statistical measures which can be used to describe the signal. Such statistical measures might supplement or replace maximum-likelihood estimates derived from correlation operations.

The previous discussion assumed only random errors which could be recognized by their statistical differences from the signal. In the investigation of advanced data detection, we must consider systematic errors which are statistically identical to the signal. Such errors cannot be eliminated by the processes described previously; however, some types of systematic errors may be corrected by appropriate calibration or by appropriate interpretation based on other data sources. For example, the drift of a strain gauge may be a predictable function of temperature so that corrections can be made to the data, provided only that we have an independent measure of temperature. In current practice, such manipulations occur only at the end of the analysis process and need not be considered in the design of equipment, only in the design of the experiment (*i.e.*, we must remember to include the temperature gauge).

The potential for later correction of such systematic errors may be lost by the manipulations of advanced data detection devices. Consequently, we must be aware of such potential and be prepared to analyze the transformation of systematic errors by the planned manipulations. If the transformation is such that these errors can no longer be compensated, even though they were (in principle) in the original form, we will have to include appropriate functions in the equipment to perform corrections prior to interfering manipulations.

Most easily applicable analyses in information theory assume stationary statistics. This means, in general, that we must predict the actual nature of signal and noise, including relative power levels, during the experiment. That such signal and noise levels are not constant throughout a typical experiment is immediately apparent when one looks at any telemetering record from a missile test.

In order to account for nonstationary statistics in practical data detection equipment, we may be able to include some adaptive features such as variable bandwidth based on signal-to-noise ratio. Such an approach is applicable if the statistics of signal and noise vary slowly relative to the smoothing time of the filter in use. If, however, we are after the greatest possible improvement and are attempting to consolidate data from long time intervals during which statistics change radically, we will be forced to utilize more complete analyses. Such analyses are available in the literature, but practical embodiments in equipment have yet to be demonstrated, although the equations involved can often be programmed for digital computation.

System Optimization

The foregoing technical discussion demonstrates that we have at hand the concepts and analyses required to effectively retrieve any experimental data which can be defined as available, based on fundamental physical limits of detectability. We must recognize, however, that it may not be economically feasible to recover all such data.

In the disciplines of information theory, optimum detection is taken to be that which results in a minimum output-error content (residual noise-plus-signal distortion). In the current context, we assume that certain output data and certain allowable errors are already defined. Then optimum filter and data reduction analysis will be applied to define the detectability of the desired data based on the input signal and noise. Having demonstrated basic detectability, the system optimization takes the form of minimization of the economic penalty (dollars) for the desired increase in informational entropy.

The first phase is the application of ingenuity to conceive and produce reliable and simple equipment mechanizing the required data handling functions. This equipment should be tested to demonstrate proper operation in confirmation of the appropriate theories.

The final decision remaining at this point would be: how much of the required data detection equipment should be designed for airborne environments and used prior to transmission of data on a radio link? This analysis is illustrated in Fig. 2.

The curves in Fig. 2 represent schematically the cost factors relative to the proportioning of the experiment. Each curve shows cost levels corresponding to possible choices for the placement x_0 of the data link. The cost of data transmission is a function of the given vehicle and

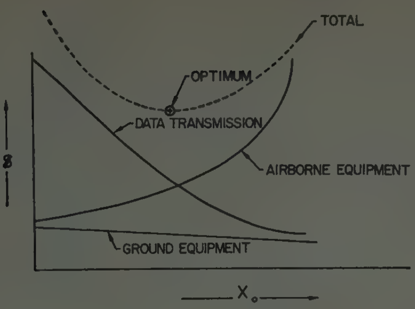

Fig. 2—Cost analysis.

its mission. Some value can be assigned in terms of dollars for each bit of data, redundant or otherwise, that is transmitted to the receiving station. The example shown implies that the available data link is capable of transmitting the entire output of the gauge. As the position of the data link is moved along the data reduction chain, the data to be transmitted are reduced and the corresponding costs are similarly reduced.

Conversely, the costs of placing the necessary data detection equipment in the vehicle environment increase, often quite rapidly. In the example shown, the vehicle does not have sufficient payload capacity to handle all the data reduction functions, and the airborne equipment costs increase asymptotically. Similar asymptotic behavior is possible for data link costs if the vehicle payload is insufficient to permit a transmitter of adequate capacity to handle the entire gauge output.

Ground-based data reduction equipment costs are usually small in comparison to airborne equipment costs which must include, in addition to development and packaging costs, a stiff tariff for the delivery of a payload pound on target (vicinity of Mars, for instance, currently 100,000 to 500,000 dollars per pound).

The sum of all the cost factors is usually a curve which displays a minimum for some choice of data link placement. The exact position of the minimum may vary considerably depending on individual judgment in assigning cost values to the various factors and the final choice of placement of the data link will usually be governed finally by the availability of specific equipments. However, an attempt at such an analysis is deemed worthwhile, if only to gain some evaluation of the potential cost of the desired data. It may even be decided that the data are not worth the economic penalty.

Analyses of the type just described are the province of the operations research specialist, but the concepts and relative values involved are of definite interest in the design of experiments. The operations research analyst would continue the process to define completely the values of each of the proposed space experiments and the relative importance of proposed new vehicle developments. (It may be better to fire more of our currently available rockets than to develop new large rockets for certain tasks.) Such an extension of the analysis is obviously beyond the scope of this paper.

Extraterrestrial Noise as a Factor in Space Communications*

ALEX G. SMITH†

Summary—Present-day refinements in communication systems make it appear that extraterrestrial noise sources may establish a fundamental limitation on long-range communications. The various cosmic and solar system radio sources are considered with respect to their intensities, spectral distributions, and temporal characteristics. The most severe forms of interference occur in the long-wavelength regions of the radio-frequency spectrum, so that the future of space communications probably lies in the perfecting of low-noise microwave systems.

INTRODUCTION

KARL Jansky first detected the cosmic radio noise in 1932 in the course of an experimental study of terrestrial sources of interference. Even after Jansky's discovery, cosmic noise was largely ignored by engineers as being of insufficient intensity to be of practical importance. Now, however, three decades later, improvements in communication systems have reached the point where extraterrestrial noise threatens to impose a fundamental limitation on man's ability to communicate over wide regions of the radio-frequency spec-

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trum. This is particularly true when one considers the problem of communication with a vehicle in the "quiet" of space.

The original cosmic noise discovered by Jansky is now firmly established as coming from a broad, diffuse region enveloping the galaxy, or Milky Way. Hundreds or thousands of small, discrete sources known as "radio stars" are superimposed on this background to contribute additional noise. The diffuse hydrogen gas which permeates vast regions of space is a radiator of microwave signals. The sun and the planet Jupiter are known to be powerful, erratic sources of radio-frequency energy, and at least two other planets may also be in this category. Even a comet has been reported as a source of radio noise. It is the purpose of the present paper to consider briefly the probable importance of each of these sources as a factor in space communications.

GALACTIC NOISE

The radio noise from the galaxy is the most widely distributed of the various extraterrestrial sources of interference. It appears to be present in some degree in all directions, but is of course most intense near the plane of the galaxy, especially in the direction of the galactic center in the constellation Sagittarius. Galactic noise has been mapped by various observers at frequencies ranging from less than one mc to 1360 mc, and the most reliable of these surveys have recently been summarized by Ko [1]. A galactic noise map in convenient slide-rule form has been prepared by Menzel [2].

Of particular interest in space communications is the level of galactic noise along the ecliptic. The majority of the more distant space missions will presumably occur in or near the plane of the ecliptic, and it is therefore the ecliptic noise level which will be present as an interfering background.

In Fig. 1 are shown curves of the ecliptic noise level for three different frequencies. The 81-mc data are from the survey of Baldwin [3], the 100-mc data are from the map of Bolton and Westfold [4], and the 600-mc curve is based on the work of Piddington and Trent [5]. The horizontal coordinate is celestial longitude, which is measured in degrees eastward along the ecliptic from the vernal equinox. The ordinate is in units of "brightness" B, where B is the flux density in watts per square meter per cps of bandwidth which would be received at the earth's surface by an antenna having a unit beamwidth of one steradian (it is assumed, of course, that B remains constant over the entire beamwidth). It has been recent practice to plot galactic noise maps in units of "brightness temperature" T, where T is the temperature in degrees Kelvin of an ideal blackbody surface which would have the same brightness as the given area of the sky. Fig. 1 has been plotted in terms of B to permit more direct comparison with the units employed in later sections. Both B and T are units which are appropriate for extended sources—that is, sources which are large relative to the beamwidth of the antenna—and they are related through the Rayleigh-Jeans approximation to Planck's Law,

$$B = \frac{2kT}{\lambda^2},$$

where k is the Boltzmann constant, 1.38×10^{-23} joules per degree K, and λ is the wavelength in meters.

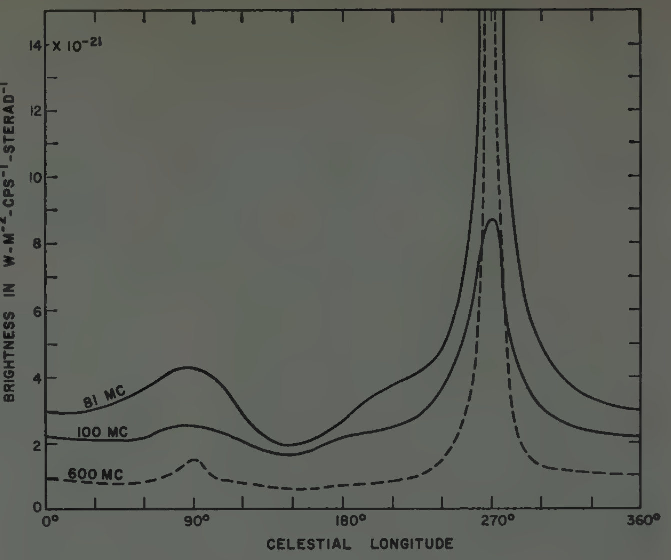


Fig. 1—Galactic noise along the ecliptic at three frequencies. The 600-mc peak extends to a height of 21 units and the 81-mc peak to 27 units.

It will be noted from Fig. 1 that at least within the given frequency range the galactic noise along the ecliptic varies only moderately (perhaps ±40 per cent) for some 300° of celestial longitude. In the remaining 60°, the ecliptic, by an unfortunate chance, passes quite close to the galactic center. Here the noise peaks abruptly, reaching, according to the surveys, values from 5 to 20 times the average level of the rest of the ecliptic. It should be remarked that the peak values are subject to considerable error because of the "smoothing" which results from attempting to resolve this sharp peak with antennas of appreciable beamwidth. In particular, the 100-mc peak is probably too low because the antenna used in the original survey had a beamwidth of 17°. By the same token, of course, if relatively low-gain antennas are employed in communications systems, the apparent galactic peak will be reduced in height but extended in longitude. Fig. 2 is an average of numerous scans through the galactic center with a 22-mc antenna at the University of Florida radio observatory. This antenna has a half-power beamwidth in the east-west direction of about 30°, and it will be seen that the galactic peak is spread over approximately 135°.

Fig. 1 indicates that the average galactic level falls with increasing frequency. The exact spectrum of the galactic noise is still in doubt because of the incompleteness of the available surveys and the uncertainties of

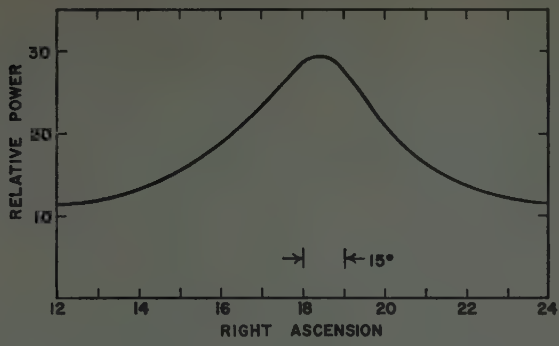


Fig. 2—22-mc galactic noise recorded in early 1959 at the University of Florida with a wide-beam antenna directed at the meridian.

ionospheric absorption at the lower frequencies. Several investigators [6] have come to the conclusion that the spectrum is of the form

$$B \propto \lambda^n,$$

where n is between 2 and 2.7 over most of the radio-frequency spectrum.

Since the galactic noise decreases with frequency while receiver noise generally increases with frequency, it is clear that it is at the lower frequencies that the noise level is apt to be established by cosmic interference. With ordinary types of receivers, the cross-over point typically occurs around 150 to 200 mc. Experience at the University of Florida radio observatory, with antenna beams ranging from 20° to 35° in width, shows that in the 18- to 30-mc region, the noise level of the best commercial communications receivers is virtually negligible compared to the galactic level at any point along the ecliptic.

It is well known that radio astronomers have detected an essentially monochromatic spectrum line due to the radiation of atomic hydrogen at a wavelength of 21 cm (this is the only noncontinuous radiation thus far confirmed). The hydrogen radiation follows a distribution in the sky roughly similar to that of the galactic noise, and reaches a maximum brightness of about 6×10^{-20} watt meter^{-2} (cps)$^{-1}$ sterad^{-1}, well in excess of the galactic brightness at that frequency. Because the hydrogen radiation is limited to a single wavelength, however, it is unimportant as a source of interference.

It naturally follows that as receivers are improved, and, in particular, as such innovations as the maser become practicable for communications, galactic noise will become a dominant factor over still wider regions of the spectrum, extending into the microwave region.

Radio Stars

Superimposed on the general background radiation due to the galaxy are numerous localized or discrete sources, generally less than 1° in extent. A list of 1936 such sources has been published by Shakeshaft, Ryle, Baldwin, Elsmore, and Thomson [7]. Since the majority of these sources cannot be identified with visible objects, they are known as "radio stars." The strongest of these sources show a tendency to occur near the plane of the galaxy.

The most intense radio star is an object known as "Cassiopeia A." At a frequency of 20 mc this source produces a flux-density at the earth's surface of 5×10^{-22} watt per square meter per cps of bandwidth [8]. (Note that in dealing with sources which are small relative to the antenna beamwidth, it is appropriate to use the concept of flux-density S rather than brightness B. The units are similar, except that S does not involve the solid angle of the beam.) At higher frequencies, the flux-density of Cassiopeia A falls off, reaching a value around 6×10^{-24} watt meter^{-2} (cps)$^{-1}$ at 10,000 mc. Most of the other strong sources show similar spectral behavior.

It is of interest to inquire to what extent such a discrete source will contribute to the received noise in a typical situation. If one assumes a frequency of 100 mc, the flux-density from Cassiopeia A is 2×10^{-22} watt meter^{-2} (cps)$^{-1}$ and the galactic brightness at the position of the radio star is 3×10^{-21} watt meter^{-2} (cps)$^{-1}$ sterad^{-1}. An antenna having an effective area A of 100 meters^2 will receive 2×10^{-20} watt (cps)$^{-1}$ from Cassiopeia A. The beam of the antenna will have a solid angle β of about 0.1 steradian ($\beta = A/\lambda^2$), and it will therefore receive 3×10^{-20} watt (cps)$^{-1}$ from the galaxy. While the contribution of the radio star is appreciable, the discrete source is of secondary importance even in this relatively favorable case of the strongest radio star and a rather weak galactic region.

The next most intense radio star is only about half as strong as Cassiopeia A, and of the remaining sources only 4 reach an intensity of 10^{-23} watt meter^{-2} (cps)$^{-1}$. In general, the noise contribution from the radio stars is minor relative to the galactic background unless extremely high-gain, narrow-beam antennas are involved.

Solar Radiation

The radio-frequency spectrum of the sun is characterized by complexity and variability. Fig. 3 is an attempt (perhaps oversimplified) to indicate the principal features of the radiation, which is often arbitrarily divided into radiation from the "quiet sun" and radiation from the "disturbed sun."

At optical and millimeter wavelengths, the radiation follows closely that expected from a blackbody at 6000°K. At longer wavelengths, however, the radiation from the quiet sun is many times more intense than would be expected from a body at that temperature. In addition, beginning at centimeter wavelengths, time-dependent increases in the radiation become conspicuous. In the region between 2 cm and 60 cm, enhancement of the general radiation over a range of about 2:1 occurs, with a periodicity of weeks or months which is

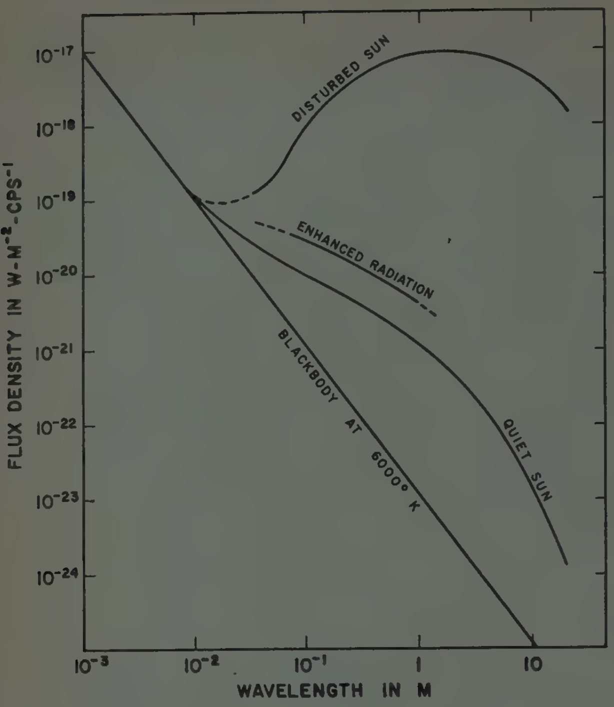


Fig. 3—Chart of representative solar noise levels, with a 6000°K blackbody curve included for comparison. The curve for the disturbed sun represents maximum probable values, but not record maxima.

associated with the area occupied by sunspots. Moderately strong outbursts of radiation in this region, lasting a few minutes, are connected with solar flares.

In the longer-wavelength region of the spectrum, beyond a meter, the general radiation level remains constant for a considerable fraction of the time, but violent disturbances are common, especially near the maximum of the sunspot cycle. So-called "noise storms," consisting of trains of bursts together with a general enhancement of the radiation, last for hours or days and show strong circular polarization. "Outbursts," lasting for a number of minutes, are associated with solar flares; they may reach intensities of millions of times the level of the quiet sun, and encompass virtually the entire radio spectrum. "Isolated bursts" of 5 or 10 seconds duration are even more frequent.

Because of the complexity of the solar radio spectrum, it is difficult to establish statistics on the percentage of the time the sun is "disturbed" at any given frequency. Certainly the disturbances are more frequent and more intense near sunspot maximum. In general, the observed level will lie between the "quiet sun" and "disturbed sun" curves of Fig. 3, although the upper curve illustrates only typical maximum intensities. At least one outburst in excess of 10^{-15} watt meter^{-2} (cps)$^{-1}$ has been recorded [9].

The sun must be ranked as second only to the galaxy as a source of interference, and it takes this secondary position only because it is a localized, rather than a distributed, source, and as such is susceptible to at least partial elimination by directional antennas. It seems certain to create severe problems in the cases of vehicles sent into its own vicinity. The planet Mercury, as seen from the earth, is never more than 28° from the sun; while the earth, as viewed from a vehicle near the orbit of Jupiter, would always be within 11° of the sun.

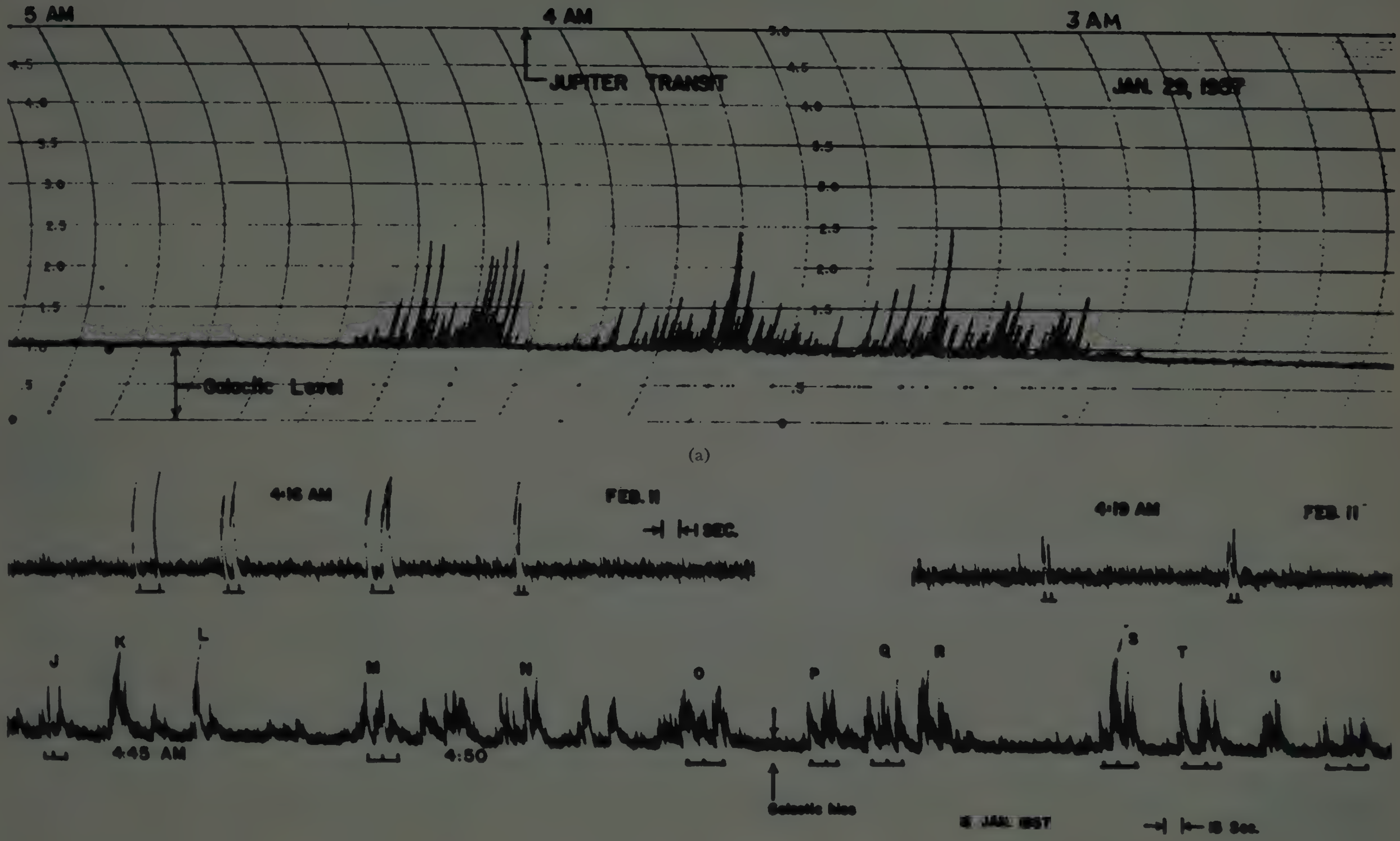
PLANETARY RADIATION

In 1955, B. Burke and K. L. Franklin, while making a cosmic noise survey at 22 mc, unexpectedly discovered that the planet Jupiter is a strong, sporadic radio source. Subsequent investigations by several observatories have indicated that Jovian outbursts may have durations ranging from a few seconds to several hours. A low-speed pen record of a typical noise storm lasting about $1\frac{1}{2}$ hours is shown in Fig. 4(a) (next page), while Fig. 4(b) shows high-speed records of the details of the Jupiter noise. These recordings were made at a frequency of 18 mc at the University of Florida radio observatory.

It will be noted that the noise occurs in bursts or groups of bursts, lasting from a fraction of a second to several seconds, with some indication of doublet or triplet patterns. The stronger bursts observed at the University of Florida at 18 mc have gone off-scale at flux-densities greater than 5×10^{-20} watt meter^{-2} (cps)$^{-1}$ while Gardner and Shain [10] have reported a peak intensity of 10^{-19} watt meter^{-2} (cps)$^{-1}$ at 19.6 mc. The observations of all workers indicate much lower flux-densities at 22 mc and 27 mc; and a search with sensitive equipment [11] has indicated that any sporadic Jovian flux present at 38 mc or 81.5 mc must be below levels of 1×10^{-24} and 3×10^{-26} watt meter^{-2} (cps)$^{-1}$, respectively. Gardner and Shain report a somewhat lower intensity at 14 mc than at 19.6 mc, but the effect of the terrestrial ionosphere at this frequency is uncertain. It thus appears that the observable sporadic Jovian spectrum lies between the terrestrial ionospheric cutoff frequency and a frequency of perhaps 30 mc.

While the Jupiter radiation is highly intermittent, statistical studies show a high degree of correlation of the outbursts with the rotational period of the planet, indicating the existence of localized sources [12]. Since these sources appear to radiate over considerably less than one-half rotation of the planet, a directional effect is involved, which several workers have attributed to a Jovian ionosphere. According to this interpretation, typical values for the critical frequency of the planet's ionosphere range from 9 to 20 mc [12].

Individual bursts appear to have a spectral width of several tenths of a megacycle, and to be circularly or elliptically polarized in the right-hand sense [13]. This polarization suggests the presence of a magnetic field, and T. D. Carr of this laboratory has used the ionospheric model mentioned above to estimate the strength of the field as 7 gauss.



(a)

(b)

Fig. 4—University of Florida records of typical Jupiter noise outbursts. The recording speed in (a) is 6 inches per hour; that in (b) is 5 mm per second. The planet takes over 2 hours to travel between the half-power points of the antenna beam, so that the starting and stopping of the radiation in (a) are actual Jovian phenomena.

Since Jupiter is second in intensity only to the sun as a localized source of interference, the frequency of occurrence of the noise is of considerable interest. Table I is a summary of the reports of several observatories of the percentage of actual observing time during which Jupiter noise was recorded. It will be seen that the 18-mc activity, *e.g.*, ranges from a high of 22 per cent to a low of 1.5 per cent. The University of Florida records indicate that, considering *both* duration and intensity, Jupiter was about 7 times as active in 1956–1957 as it has been in the subsequent two years. Whether these variations can be correlated with other phenomena, such as the sunspot cycle, remains to be established. In a number of instances during the 1957–1958 apparition, short bursts of noise were observed at the University of Florida immediately prior to sunrise, during periods when the planet seemed to be inactive. Carr has hypothesized that these events may be due to transitory focusing of low-level Jovian radiation by a terrestrial ionosphere gradient [13], in which case it may be that the planet emits weakly much or all of the time.

TABLE I

Percentage of Observing Time During Which Jupiter Radiation Was Detected, 1955 to 1959

Source of Data	Year	Frequency	Percentage of Time Jupiter Active
Gardner and Shain [10]	1955–56	19.6 mc	5 per cent
Gallet [22]	1956	18	8.5
Gallet	1956	20	16.9
University of Florida	1956–57	18	22
Gallet	1957	18	6.1
Gallet	1957	20	10.6
University of Florida	1957–58	18	1.5
University of Florida	1957–58	22.2	0.8
University of Florida	1957–58	27.6	0
University of Florida	1959	18	4
University of Florida	1959	22.2	1.5
University of Florida	1959	27.6	0.1

Extensive monitoring of Saturn at 18 and 22 mc by the University of Florida observatory has resulted in the recording of several disturbances of possible Saturnian origin during the 1957 and 1958 apparitions. These events resembled weak bursts of Jupiter noise, lasting only a few minutes and reaching a maximum intensity of 6×10^{-21} watt meter^{-2} (cps)$^{-1}$. Allowing for the greater distance of Saturn, these bursts are comparable to the stronger Jovian outbursts. Smith at Yale [14] has also recorded possible Saturn noises at similar frequencies. At the present time, the low altitude of Saturn as seen from the United States, as well as its location in the noise peak near the galactic center (Figs. 1 and 2), impose observational difficulties which have prevented the results from being conclusive.

In 1956, Kraus reported bursts from Venus which reached an intensity of 9×10^{-22} watt meter^{-2} (cps)$^{-1}$ on 26.7 mc. The bursts consisted of short pulses of much less than a second's duration, and of longer pulses which lasted up to several seconds and had a spectral width of at least 2 mc [15]. Several months of monitoring Venus at 27.6 mc at the University of Florida observatory in 1958 failed to recover the radiation [13]. It is possible that the Venusian activity observed by Kraus has suffered a temporal decline similar to that of Jupiter.

The prominent Comet Arend-Roland (or 1956h) was reported as a radio source by Kraus [16] and by Coutrez, Hunaerts, and Koeckelenbergh [17]. Coutrez, *et al.* measured a peak flux-density of 8×10^{-23} watt meter^{-2} (cps)$^{-1}$ at 600 mc when the comet was at a distance of 0.7 astronomical unit, while the maximum intensity recorded by Kraus at 27.6 mc was 5×10^{-22} watt meter^{-2} (cps)$^{-1}$. Conspicuous comets are, of course, of such infrequent occurrence as to be relatively unimportant as potential sources of noise.

Radio-frequency thermal radiation has been measured for the moon and for the planets Venus, Mars, Jupiter and Saturn. For the moon, at a frequency of 35 kmc, the received thermal flux density [18] approximates 10^{-21} watt meter^{-2} (cps)$^{-1}$; outside of the microwave region the thermal flux is presumably immeasurably small. Planetary thermal fluxes are similarly detectable only at very short wavelengths. The largest reported is that of Venus [19], 10^{-24} watt meter^{-2} (cps)$^{-1}$ at 9500 mc, while the smallest is a value of 4×10^{-26} for Saturn at 8000 mc [20]. At present, these fluxes are obviously of only minor importance as sources of interference, requiring integration times for their bare detection which would be impracticably long for communication systems.

Quite recently, observers at the National Radio Astronomy Observatory, the California Institute of Technology, and the Naval Research Laboratory have detected anomalously strong microwave radiation from Jupiter at wavelengths ranging from 3 cm to 68 cm [21]. Although the radiation is intense enough to indicate unrealistically high temperatures if it is interpreted as thermal radiation (some of the "temperatures" run as high as 70,000°!), the fluxes are still of the order of only 10^{-25} watt meter^{-2} (cps)$^{-1}$. Drake has suggested that the radiation may be due to a Jovian Van Allen Belt.

Conclusion

Table II is an attempt to list maximum values of the interference which may be produced at position of the earth by the various sources discussed above. For the solar system sources, the inverse square law must of course be applied to find the intensities at other points in space.

The galactic noise is the most ubiquitous form of interference, and, except for the case of communications taking place within a small angular substense from the sun, it is probably the most serious. The planet Jupiter is an extremely noisy body which is certain to cause problems with communications systems in its vicinity

TABLE II

APPROXIMATE MAXIMUM INTENSITIES OF VARIOUS SOURCES*

Source	Maximum Observed Intensity	Wavelength Region of Greatest Intensity	Temporal Characteristics
Galactic noise	10^{-19} watt meter^{-2} (cps)$^{-1}$ sterad^{-1}	15 meters	Constant
Hydrogen line	6×10^{-20}	21 cm	Constant
Sun	10^{-15} watt meter^{-2} (cps)$^{-1}$	1–10 meters	Sporadic
Jupiter	10^{-19}	15 meters	Sporadic
Saturn (?)	6×10^{-21}	15 meters (?)	Sporadic
Venus (?)	10^{-21}	10 meters (?)	Sporadic
Moon (thermal)	10^{-21}	*M* meters	Constant
Radio stars	5×10^{-22}	15 meters	Constant
Comet	5×10^{-22}	10 meters	Variable
Planets (thermal)	10^{-24}	*M* meters	Constant

* Data do not extend below 20 mc, since ionospheric absorption introduces serious uncertainties at that point.

(although, by the same token, the Jovian radio outbursts might be used to actuate homing devices). Saturn and Venus may also belong in the category of sporadic radio-frequency emitters, although the evidence in these cases is not as yet clear-cut.

Examination of the spectral characteristics of the various sources indicates that, except for thermal emission, interference is most serious at the long-wavelength end of the radio-frequency spectrum. This would suggest that the future of long-range communication systems lies in the development of low-noise microwave devices such as the traveling-wave tube and the maser. Because of ionospheric absorption, knowledge of the spectra of extraterrestrial sources at frequencies below 20 mc is incomplete and inaccurate. The extension of measurements into this region is one of the exciting possibilities associated with lunar or space observatories.

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Space Communications Requirements of the Department of Defense*

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Summary—**Military requirements for space communications are described in terms of general needs rather than specific applications. Communications to meet these needs may be divided into communications to, from, and among objects in space. The functions of these communications are command, tracking, and data transmission, the major considerations affecting each function being briefly described.**

Communications by means of objects in space, which is of most interest here, is comprised of systems utilizing either passive or active relay objects. The latter systems may be delayed retransmission types or real time repeaters, both of which may have military application.

THERE is enough uncertainty as to what is meant by a military requirement to suggest that any discussion of the the subject should begin with at least a perfunctory explanation. There are, of course, formal, stated military requirements generated by the Military Departments individually, or jointly through the Joint Chiefs of Staff. In general, these requirements will be in areas where the need for a weapons system or components is clearly defined and the technology that supports it is well advanced. This type of requirement may be carried so far as to attempt to describe the weapons system to meet the need, rather than just the need itself. However, needs or requirements, in the broad sense, can exist without the technology to support them, and technology can exist without the need for its application to a military use. There are, accordingly, areas where the technology to meet the need is new or perhaps does not exist. This latter situation is true for much of space communications, where application of the latest advances in orientation control systems, low-noise receivers, modulation techniques, and efficient converters of power to radio frequencies are required by the great cost of putting payloads into space. These applications will grow and become better defined as a result of the continuous relating of an improved appreciation of these needs and the developing technology.

Space communications is defined as communications with and among objects in space. Communications by means of objects in space may be considered to be an application of the techniques of space communications.

In regard to communications with objects in space, the communications link to the object is often easier to achieve than the link from it, because the information rate required is usually considerably less, since it is limited to a few discrete preselected commands. This is true because the space object is usually the source of information obtained by measuring devices in the object. However, even for those missions where the information rate can be considered high, as in, for instance, the outgoing leg of a television relay satellite, the average power and antenna aperture needed for the work are not so limited by weight considerations, highly directive antennas being practical on the ground. The problems that do exist for the outgoing link still center around the space-borne equipment, involving, in addition to the overwhelming problem of reliability in a difficult environment, resistance to interference and, more specifically, linearity of signal channels.

The incoming leg of a communications link with a space vehicle will be more difficult, primarily becauee the power available for the purpose will be restricted by the very high launching costs per unit weight, and high reliability will be required for high-power components. Antenna aperture will be limited both by the cost and also by considerations of constructing and automatically and reliably orienting such apertures in space. In addition, for some applications, the need for illumination of a large area on the earth rather than the single addressee characteristic of the ground-to-space leg may restrict the antenna directivity and consequently the aperture and gain available.

Communications among objects in space unfortunately has all the problems of ground-to-space and space-to-ground, plus additional difficulties associated with initial and continued orientation of antennas, which are relieved only slightly by the availability of a wider frequency spectrum for the purpose (below the VHF and above the SHF bands). The technical considerations of communications ground-to-space, space-to-space, and space-to-ground have been quite concisely discussed in an article by Rechtin.[1]

Military requirements for communications with objects in space involve the functions of command, tracking and data transmission. It is well to recall here that the purpose of communications is to transmit information and that information consists of change, if change is defined to include more precise determination of a steady-state condition. Accordingly, the purpose of commands is to achieve a change; for example, to change the path of a space vehicle, perhaps to achieve a rendezvous with a previously launched satellite; to direct recovery of a data package at a particular location; to start or stop a data transmitter; and to perform a se-

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quence of tasks or experiments that cannot be preprogrammed but may be dependent on earlier data gathered by the space vehicle. Commands are accordingly required for satellite interception, recovery of weather photographs, readout of stored data, and directing the acquisition of particular data that have been found to be of interest.

Tracking of space vehicles consists of measuring range, range rate, angles and angular rates of a space vehicle from an observation point. The space vehicle's path comprises information, or more correctly, what is not already known about its path comprises information. This information is communicated by the change (over what can be predicted) of length and direction of an electromagnetic path.

Tracking of space vehicles for military purposes is a difficult subject to discuss briefly. In general, it may be said that any vehicle must be tracked well enough so that it can be found, unambiguously, at any future date. The accuracy of tracking and, correspondingly, the accuracy of prediction of its path required for this purpose, varies with the time lapse expected before the next observation, and the number of other objects in space and the similarity of their paths. The frequency of observation is largely determined by unpredictable perturbing influences, such as uncertainties in air drag; of course, more frequent observations improve the measurement accuracy. This sets a minimum value for comprehensive knowledge of vehicle paths. For a particular military objective the accuracy required may greatly exceed the minimum, as in navigation, mapping, and geodesy. It may also be less than this so-called minimum, when a comprehensive knowledge of all space objects for a particular mission is of no consequence. An example of this is the delayed repeater communications satellite, where at remote locations the satellite's orbit need only be known well enough to acquire it in a moderately broad beam on the next usable orbital pass.

Data transmission from military space vehicles will be similar to that for nonmilitary use with regard to the variation in need for amounts of information to be transmitted. Most missions will require relatively low amounts and rates of transmission. Others, such as meteorology, are limited by the information storage capacity and, more basically, by the power that can be provided in the space vehicle. Differences between the military missions and the nonmilitary are characterized for the military by an additional need for security, both in the characteristics of the data sent and in its availability for readout, and by added resistance to interference.

Communication by means of space vehicles has important potentialities to meet military needs. Stated broadly, communication beyond line-of-sight but using what were formerly considered "line-of-sight" frequencies has rapidly become possible by artificially induced means. Long distance point-to-point communications become limited by the power that can be produced in (or reflected from) a satellite, rather than being limited by the usable bandwidth that is available in the HF and lower spectrums or the severe signal attenuations inherent in natural tropospheric and ionospheric scatter circuits.

Two approaches to communications by space vehicles are possible: the use of either active or passive space objects as relay points. These objects may utilize frequencies from 20 mc to above 50 kmc and, in theory, optical frequencies corresponding to the atmospheric "windows."

Passive space relays are attractive by reason of their simplicity, reliability and lack of intermodulation problems. They are unattractive because of their intimate relationship to the "fourth power" law that makes the received power a function of the fourth power of the distance from the transmitter to the space object to the receiver. Distances in space are inconveniently large for this purpose. Passive objects may be metallic balloons, dipoles, assorted configurations of flat surfaces, and miscellaneous debris such as burnt-out final stages of launching vehicles and nonradiating satellites.

Active space objects for communications are perhaps of more military interest than the passive objects, if the reliability problems can be solved. These active types can be divided into two categories, delayed relay and real time relay. In the first type, information via a ground-to-space link is stored in a satellite flying at a relatively low altitude, physically transported to a new location, and the stored information is read out at this location. The well-known example of this technique is the SCORE Project. This type has the advantages of short terminal transmission paths (a few hundred miles), low demands on the launching vehicle's performance or, conversely, higher potential power in the space equipment and consequently a higher data transmission rate. It has the disadvantages of an appreciable time delay due to the satellite travel time plus additional delays because of synchronization incompatibility. These latter delays are a function of the orbit period and the earth's rotation period for particular inclinations of orbit and site locations. For military purposes there are some possible advantages in this type from a security point of view, because of the limited areas from which a low-flying satellite may be seen.

Low flying satellites do not necessarily have to use this relay technique. An interesting possibility is to use a number of satellites with a prescribed orbital phase relationship to each other whereby, in cooperation with a corresponding series of ground stations, a multiple hop HF communications system in real time may be simulated at VHF and UHF.

A more intellectually satisfying approach is that of the single vehicle real time repeater. In order to be seen from widely separated distances on earth a single vehicle must, of course, be at a high altitude. This criterion is met at about 8000 miles altitude. However, the energy required to impart this circular orbit altitude corresponds to that required to escape from the earth's gravitational field and, in addition, requires a number of satellites in a pattern to provide continuous point-to-point communications. A satellite at about 19,000 nautical miles altitude, variously called a 24-Hour, Stationary or Synchronous Satellite, becomes interesting because only one satellite is required for continuous, point-to-point (or point-to-area) communications for nearly a hemisphere of the earth's surface. However, this requires considerably more energy for injection than even an escape mission requires, so that for a given vehicle a lesser payload is possible than for satellites at lower orbits. At the same time the power requirements have gone up in proportion to the square of the additional distance or, more realistically, the bandwidth that can be transmitted has been reduced in proportion to the square of the distance.

These considerations must be balanced against each other and against many others in any selection of a space communications system to meet military needs. The low-flying delayed relay satellites appear to have promise of handling a great amount of the routine communications traffic associated with logistics and personnel operations where moderate time delays are not important and may actually be less than that due to normal HF traffic backlogging. This will relieve the HF circuits for greater reliability in higher priority messages. The higher altitude satellites can be used to provide reliable real time communications where the bandwidth requirements are compatible. As launching vehicle efficiencies improve, in terms of pounds of payload in orbit per dollar, and as reliability and useful life problems are solved, satellites will become increasingly important for military communications.

Communication Satellites*

DONALD L. JACOBY†

Summary—**This paper is intended to provide a review of communication satellites starting with their comparatively brief history and continuing through present developments and future concepts and the probable impact on global communications.**

The advantages and disadvantages are considered for various types of passive and active devices, for various orbits, and from the standpoint of both military and commercial applications. Consideration of all major technical and economic factors leads to the conclusion that a system of 3 or 4 repeater satellites in 24-hour orbits will best meet future world-wide communication requirements. A number of system concepts are described, including time-synchronous and frequency-sharing systems.

Important aspects of communication satellite design are discussed including electronics, antennas, power supply, structure, attitude and position stabilization, command control, and telemetering. Selection of frequency is shown to be not critical over a wide range. Choice of design parameters such as power output, antenna gain, stabilization accuracy, etc., is largely dependent on the desired capacity of the communication system and the available satellite payload. Some typical parameters for a high capacity system are presented.

Reliability is an over-riding consideration in satellite design and it is proposed to achieve desired life by proper choice of components, by use of redundancy, and by minimization of satellite requirements at the expense of the ground equipment.

Based on published information on ARPA-NASA programs and booster capabilities, some estimates are made of the rate of progress which can be expected in communication satellites, leading to their eventual use in military and commercial communications. Wide application is foreseen because of increasing communication requirements which will saturate existing facilities.

INTRODUCTION

THE next ten years will see a revolution in world-wide communications with the advent of the communication satellite. In this age of jet planes, nuclear power, and moon rockets, our communication facilities have simply become outmoded. These facilities have served us well and will continue to do so in the future, but they are not good enough. They will not handle the greatly increased volume of communications. They will not support world-wide television which could contribute to improving international understanding. Present communications are too easily disrupted. A slight increase in sunspot activity can cause all radio communication to Europe to be blacked out. A fishing trawler may accidentally cut the North Atlantic telephone cable. While these events present difficult problems to civilian communications, the implications of even temporarily being cut off from the rest of the world are far more serious from a military standpoint.

During the past two years, communication satellites have attracted considerable interest within both industry and government. There has been a successful launching of a communications satellite (Project SCORE), and development programs have been initiated both within the civilian space agency and the Department of Defense. Rather than discuss any specific programs, an attempt at an underlying philosophy of communication satellites will be presented, starting with the objectives.

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OBJECTIVES

What should be some of our major objectives in developing a system of communication satellites? First of all, we are looking for *reliable* communications which provide uninterrupted service over long periods. The system should have *high capacity* with capability for handling large volumes of all types of traffic. It should be sufficiently *flexible* to serve the maximum number of potential users. There should be *minimum delay* in transmission. Last but not least, it should be *economical* in terms of the cost of the services rendered. Each of these factors will be discussed in more detail, and the communication satellite will be compared with some of our existing facilities.

Reliability

There are two types of reliability of interest to us. The first of these is propagation reliability. The high-frequency band has always been subject to the vagaries of the ionospheric layers which surround the earth. Thus, only a portion of the HF band is actually useable at any given time over a particular path. In addition, multipath effects seriously limit the amount of information which can be transmitted over a given channel. Added to these limitations are the blackouts which may result from ionospheric disturbances caused by sunspot activity. (As indicated in the *New York Times* last March, similar disturbances can be introduced by high-altitude nuclear explosions.) One is forced to the conclusion that HF radio via the ionosphere is not highly satisfactory as a propagation medium.

By contrast, a communication satellite of the active repeater type employing line-of-sight transmission at microwave frequencies would be extremely reliable from a propagation standpoint. However, the communication satellite introduces a different type of reliability problem, that of reliable unattended operation for long periods while in orbit. Some experts doubt that a reliability sufficient to make a communication satellite economically feasible can be achieved. It is believed that a reliable communication satellite can be developed with a minimum guaranteed life of one year based on the following assumptions: there is major effort to develop components of proven reliability for this application; all components are operated well within their ratings; the satellite design provides adequate protection during the launching and while in the space environment; and, finally, adequate use is made of redundancy to further increase the probability of successful operation. Despite the formidable problems, the satellite does enjoy some advantages. Unlike ground equipment, particularly military, which must withstand extremes of operating temperature, humidity, atmospheric pressure, and mechanical shock, the electronic equipment in the satellite will be provided, once in orbit, with an almost fixed environment devoid of the factors that generally contribute to reducing component life.

High Capacity

Through the years we have been dependent, almost exclusively, on high frequencies in the band 5 to 20 megacycles for our long range global communications. These frequencies are shared among all countries and must support both military and civilian applications. The narrow range of frequencies, and the propagation characteristics discussed previously, seriously limit the total communication capacity.

It is not surprising that there is always great interest in new techniques which promise to open additional areas of the frequency spectrum to long-range communications. Recent examples are ionospheric and tropospheric scatter propagation. The communication satellite will open up the complete range of frequencies to 10,000 megacycles for long range communications, thus providing nearly 1000 times the spectrum available in the HF band.

Flexibility

One of the requirements is to provide sufficient flexibility in our system so that new or changing demands around the world can be satisfied without major overhaul or replacement of facilities. A major disadvantage of the submarine cable, for instance, is its lack of flexibility, being strictly a point-to-point fixed-plant type of facility. On the other hand, a system of communication satellites in 24-hour equatorial orbits, by providing wide bandwidths and essentially global coverage, places a minimum of restraint on the number and location of ground stations served and the volume of communication furnished to each.

Minimum Delay

Another of our objectives is to speed up our communications. All too frequently urgent messages are delayed because our facilities are congested or because propagation conditions are poor. Recently, a television program was seen in the United States only a few hours after it was broadcast in England and this was considered a major achievement. Some of the advantages of communication satellites previously discussed, such as the wide bandwidth, will not only make possible worldwide television broadcasting in real-time, but will reduce delays in all types of communications.

Economics

If one accepts the fact that communication satellites will improve global communications, it is still reasonable to ask how much they will cost. Here we find that communication satellites have some unique characteristics. In the first place, the quantity of satellites required is not sufficient to warrant production tooling, and, therefore, they will always be expensive by comparison with mass-produced equipment. Of far greater significance, however, is the cost of the launching vehicles, which may run 5 to 10 million dollars or more depending on the satellite weight and the orbit. The present state of the art requires perhaps three launchings to insure one successful satellite, but missile experts hope that this ratio can be improved in the future. If the most pessimistic figure is taken, it could cost as much as 90 million dollars for the vehicles alone for a system of 3 satellites in 24-hour orbit. Once in orbit, a communication satellite has no operating and maintenance costs in the usual sense. However we must plan on periodically replacing satellites which have failed. It is easy to see that, on a long term basis, the cost of a communication satellite system is thus intimately tied to the reliability of the satellite, and a few million dollars to improve reliability during development would save many times that amount during each year of operation.

Survey of Types of Communication Satellites

Thorough review of all factors has led to the conclusion that an active repeater type of satellite in a 24-hour equatorial orbit offers the most promise for advancing global communications. However, in view of the reliability problems and the anticipated costs of a communication satellite program, it would be well to briefly review the pros and cons of alternative approaches which are under development or have been proposed.

Passive Reflector

A concept which has attracted considerable interest is that of the passive reflector satellite as described recently.[1] Some of the advantages of such a system are its inherent reliability and the possibility of being shared by a large number of users operating over a wide range of frequencies. However, a typical system operating between two locations 2000 miles apart would require a total of 24 100-foot balloons in randomly spaced orbits at 3000 miles altitude for an outage time of 1 per cent. In addition, to support meaningful communication bandwidths, extremely large transmitter powers, low-noise temperature receiving systems, and large tracking antennas with beamwidths of the order of 0.05° are required. Substantially greater numbers of satellites would be required to provide longer range or wider coverage than the examples cited and, therefore, they do not appear to be an economical solution to providing truly global communications.

The passive reflector satellites would be more attractive if they could be stabilized in attitude, permitting use of more efficient reflecting surfaces, or if the satellite orbits could be synchronized so as to reduce the number of satellites required. However, the satellites then cease to be "passive" and reliability is no longer inherent in the system.

Delayed Repeater Satellite

This concept envisons a satellite in low-altitude (300 to 2000 miles) orbit providing communications between a number of ground stations. Each ground station provides facilities for storing communication traffic destined for other stations and transmitting this traffic to the satellite during the time the satellite is in view. Simultaneously, the ground station will receive from the satellite traffic from other ground stations which have been stored in the satellite. The satellite is equipped with recorders and appropriate control circuits which can be commanded from the ground or perhaps operate on a prearranged program.

A satellite in equatorial orbit can provide communications to stations in the equatorial zone on each satellite orbit. In orbits other than equatorial, a satellite will serve a larger area but will in general serve a particular station less frequently. The higher the altitude of the satellite, the longer is the time in view for each ground station and the larger is the area served.

The delayed repeater satellite appears to be an economical method of providing communications where delays of 2 to 3 hours are tolerable. It is most efficient when operating in the region of the equator. It is not intended to provide global coverage or communication of a high priority nature.

Communication Satellite in 24-Hour Orbit —System Design

This concept is illustrated in Fig. 1. Three satellites at an altitude of 22,300 miles and equally spaced over the equator circle the earth in 24-hour equatorial orbit. Because of the earth's rotation, they appear to remain

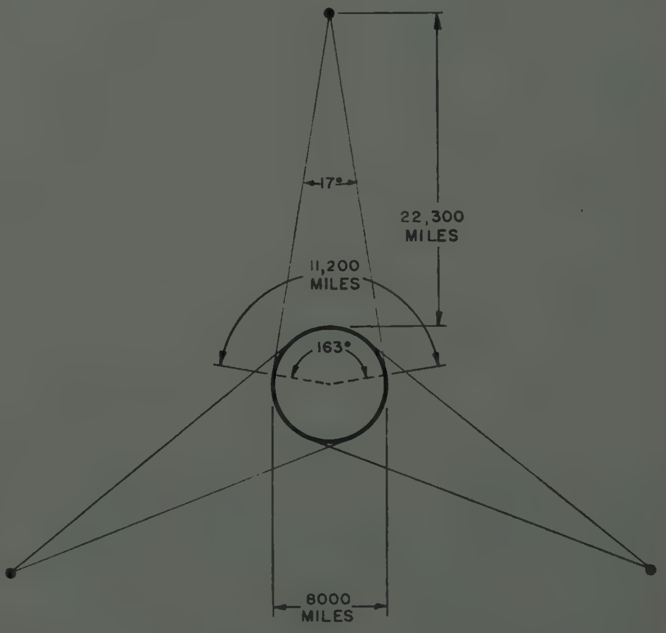

Fig. 1—Concept of three communication satellites in 24-hour equatorial orbit, showing geometrical relationships in the equatorial plane.

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[1] J. R. Pierce and R. Kompfner, "Transoceanic communications by means of satellites," PROC. IRE, vol. 47, pp. 372–380; March, 1959.

fixed to an observer on the earth. Such a satellite system will provide complete global coverage with the exception of the extreme polar areas. Each satellite will be equipped to receive, amplify and retransmit signals originating at stations within its area of coverage and intended for other stations in the same area. Switching between areas could be accomplished by direct relay between satellites, but since this might prove difficult to control, a preferred method is to locate ground repeaters in the areas visible to 2 satellites, as shown in Fig. 2.

Within the concept described there are at least three approaches to system design. In the first of these [see Fig. 3(a)] and the most straightforward, two frequency bands are employed, one for transmissions from satellite to ground and the other for transmission from ground to satellite. These frequency bands are subdivided, not necessarily equally, among the ground stations and the inter-area repeaters.

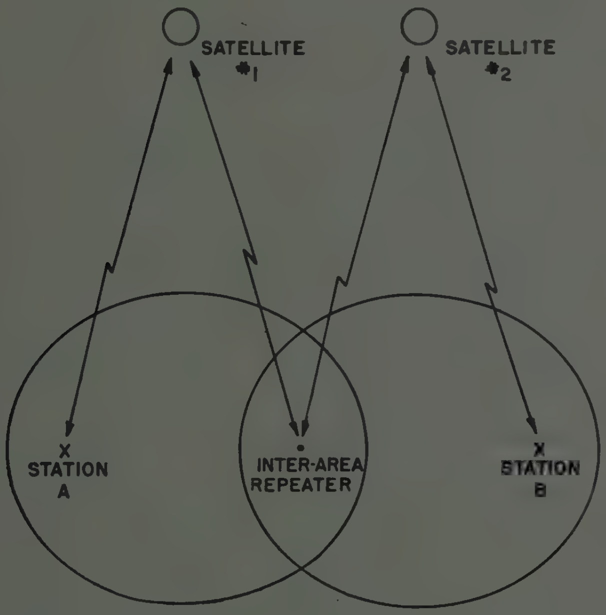

Fig. 2—Method of switching between areas in a 24-hour satellite communication system.

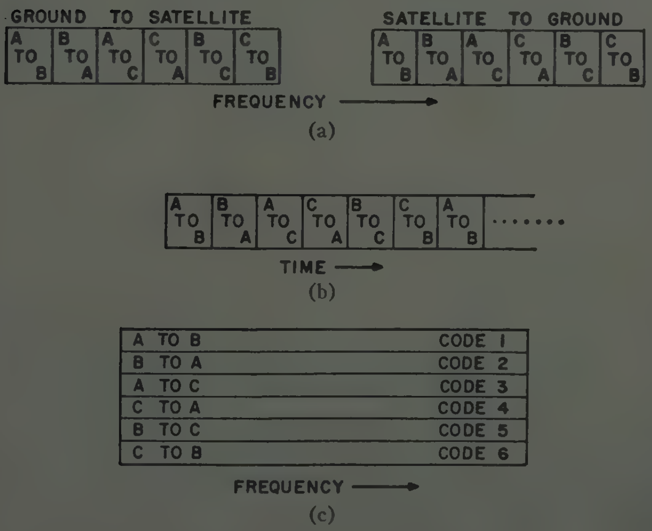

Fig. 3—System design approaches in a 24-hour satellite communication system.

In the second approach [see Fig. 3(b)] all stations would have common frequencies, but would transmit and receive at different times. This system entails greater complexity than the first because of the need for maintaining precise synchronism between stations and the probable variations in delay time in the transmission path due to fluctuations in satellite position.

A third approach would again have all stations sharing common frequencies, but with all stations transmitting and receiving simultaneously. Each signal would be furnished with an identifying code. Separation of signals at receiving stations would be by correlation detection techniques [Fig. 3(c)].

Some discussion of frequencies to be employed for these systems seems appropriate at this point. Frequencies below a few hundred megacycles should be avoided because of ionosphere effects. Frequencies above 10,000 megacycles are subject to attenuation in the atmosphere due to oxygen, water vapor, and rain in varying degrees, the attenuation being most pronounced for grazing paths and less significant for paths normal to the earth's surface. Having established rough upper and lower limits on frequency, the basic formula for transmission in free space will be examined.

$$P_R = P_T + G_T + G_R - 37 - 20 \log f - 20 \log d$$

where

P_T = transmitted power in dbw,
P_R = received power in dbw,
G_T = transmitting antenna gain in db (relation to isotropic),
G_R = receiving antenna gain in db (relation to isotropic),
d = distance in miles,
f = frequency in megacycles.

If we now decide that the satellite antenna should have a fixed beamwidth such that it exactly illuminates the earth, and if we further establish a practical limit on the size of the ground antenna, only one of the antenna gains will be a function of frequency and P_R will then be independent of frequency. Thus, the actual choice of frequency boils down to availability of reliable components, such as tubes having necessary power output, bandwidth, efficiency, and considerations of mutual interference with other users of UHF and SHF frequencies.

Noise is not expected to be a serious problem in the ground-to-satellite path since high-power transmitters may be employed. However, satellite transmitter power may be limited by payload restrictions in which case use of parametric amplifiers or masers may be required to minimize the effective noise temperature of the ground receiver.

Table I lists typical system parameters for a hypothetical communication satellite system handling 12 PCM voice channels between a pair of ground stations.

TABLE I

Frequency	Satellite	Ground
Transmitter Power	2000 megacycles	2000 megacycles
Antenna Gain	18 decibels	49 decibels
Receiver Bandwidth	1 megacycle	1 megacycle
Receiver Noise Figure	10 decibels	3 decibels
Carrier to Noise	37 decibels	24 decibels
Channel S/N		60 decibels

SATELLITE DESIGN

Some aspects of communication satellite design, which are of particular significance if we are to meet our over-all system objectives, will now be briefly discussed. More detailed information is available in a recent paper.[2]

Weight Limitation

For the next few years we will be severely limited in the payload that can be placed in orbit by available boosters. These capabilities can be expected to increase and within 10 years we should be able to place payloads of several thousand pounds into 24-hour orbit. However, it is generally true that the heavier the payload, the greater will be the cost of the booster to place it in a given orbit. Therefore, it is important to design for minimum weight by use of materials and structures with maximum strength-to-weight ratio, by employing the most efficient power sources, and by minimizing power requirements.

Stabilization

Attitude stabilization of the satellite is essential to positive control of the satellite orbit and is also a major factor in weight reduction. It permits use of a directional antenna with consequent reduction in power requirements. It also permits the incorporation of solar cell platforms that rotate to always face the sun, thus improving power supply efficiency by a factor of 6. Attitude stabilization may be achieved by means of sun and star trackers, horizon sensors, or radio signals from the ground, controlling a system of gas nozzles.

Antennas

The satellite antenna may be a single fixed parabola. Linear polarization is satisfactory, provided that the ground antenna is circularly polarized. In the case of a 24-hour satellite, a beamwidth of 20° will illuminate the earth with some margin provided for attitude variations. Higher gain antennas may be employed to illuminate specific areas or increase system margin but this either places more severe requirements on attitude stabilization or requires some form of steerable or electronically scanned antenna.

Power Supply

As discussed previously the most efficient power supply in the foreseeable future is a solar cell platform which rotates so as to always face the sun. Present solar cells have an efficiency of 10 per cent and will provide approximately 10 watts per square foot of platform area. Some improvement in efficiency, perhaps to 13 per cent, is foreseen in the future. A battery supply should be added in parallel with the solar cell supply if operation is desired during periods when the satellite is in the earth's shadow. This amounts to less than 1 per cent of the time for a satellite in 24-hour orbit.

Communication Electronics

The major emphasis in the design of the communication electronics must be on reliability. We must select only components of proven reliability; we must operate them well within their ratings; and components must be protected from damage by micrometeorites and radiation. Present microwave tubes are not capable of operating continuously for a year or more for this application and extensive research and development is required in this area. Reliable components will be supplemented by prudent use of redundancy to insure maximum useful life of the satellite. A likely circuit configuration is shown in Fig. 4. This is basically a heterodyne repeater, with traveling-wave tubes a logical choice for the wide-band amplifier stages.

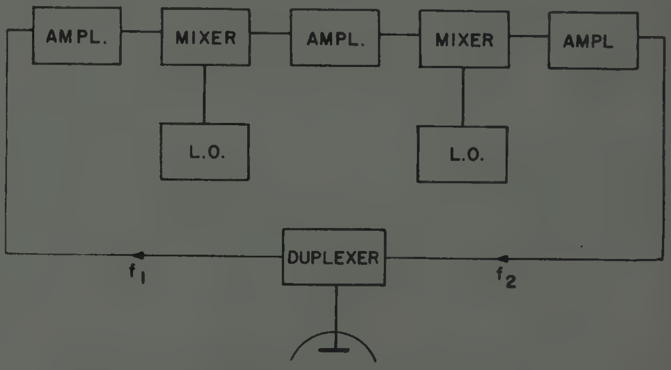

Fig. 4—Basic design of communication satellite repeater.

Command System

The satellite will undoubtedly require some degree of control from the ground through a command link. Such a link could provide control of satellite position as well as commands to activate or deactivate portions of the satellite communications equipment and other functions. Commands would be in coded form to minimize the possibility of interference.

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[2] J. E. Bartow, G. N. Krassner, and R. C. Riehs, "Design considerations for space communication," 1959 IRE NATIONAL CONVENTION RECORD, pt. 8, pp. 154–166.

Telemetry System

There is also a requirement for a telemetry link from the satellite to the ground to provide information vital to operation of the communication satellite system. The telemetry system will monitor the performance of the communication equipment, the power supply, and the attitude stabilization system and will provide acknowledgment of commands and verification that they have been acted upon. During the research and development phase, the telemetry link serves the additional important function of checking the adequacy of the design, the information derived serving as a basis for design modifications in later satellites.

Orbit Correction

Ideally, a satellite in a 24-hour equatorial circular orbit will remain over a fixed point on the earth. In practice it is anticipated that initial inaccuracies together with effects of the moon's gravity, solar radiation pressure, etc., will bring about perturbations in the orbit requiring periodic correction during the useful life of the satellite. In order that the satellite orbit may be properly controlled from the ground, it is likely that a tracking beacon will be incorporated in the satellite. Tracking will be required at a number of ground locations unless the tracking beacon is of a transponder type, supplying range in addition to angle information. Doppler measurements are useless in tracking a 24-hour satellite because of the small magnitude and unpredictable variations of differential velocity. Information from the ground tracking stations will be fed to a computer which will determine the direction and timing of the corrective thrusts which must be applied to the satellite. It appears feasible to maintain a satellite within 1 or 2 degrees of its desired location by these techniques.

The command, telemetry, and tracking systems may operate close to the communication band and share common antennas. However, they do not require wide bandwidths and could operate at considerably lower frequencies. The decision may depend on whether tracking and position control are performed at the communication ground station or at separate locations.

Structure

As discussed previously, there will be a maximum weight for the satellite. There will also be maximum dimensions for the satellite during the launching period as dictated by the final stage rocket configuration. Hardware which will project from the main body of the satellite, such as the solar cell platform and the antenna, will be in a collapsed position during launching, and extended only after separation from the last stage. The satellite must be designed to withstand the conditions of vibration and shock during launching and to have a maximum degree of dynamic stability in orbit.

Thermal Design

The satellite must provide a satisfactory thermal environment for the electronics and other components. The power dissipated in the form of heat by the electronic equipment must be brought to the outer surface whence it can be radiated. The outer surface must be designed, using special coatings as required, to strike a balance between radiation and absorption such as to maintain internal and external temperature within desired tolerances. In the particular case of the solar cells, it is essential that close thermal control be maintained, since their efficiency drops off rapidly at higher temperatures. Thermal storage devices may be required to prevent wide variations in thermal conditions as the satellite traverses its orbit.

CONCLUSION

Serious shortcomings exist in our present global communication facilities. Communication satellites promise greater capacity, wider coverage, more flexibility, and fewer delays than present transmission systems. The operating cost of communication satellites is intimately tied to the problem of reliability. With sufficient emphasis on the reliability aspects, satellites will assume the prominent role in communications for which they appear destined.

Interference and Channel Allocation Problems Associated with Orbiting Satellite Communication Relays*

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AND H. F. MEYER†, SENIOR MEMBER, IRE

***Summary*—The use of active and passive relays is considered for application to long-distance global trunk communication, with particular emphasis on economic utilization of the media from the traffic capacity viewpoint. Active and passive techniques are compared with regard to power requirements, coverage, mutual interference, and ability to exploit wide-band modulation systems to reduce interference. With the emphasis on microwave transmission, narrow beamwidths, and a multiplicity of relays, channel allocations of the future will necessarily be determined by spatial considerations, in addition to specifying frequency bands. Examples with typical global paths and assumed active and passive systems are illustrated. It is concluded that, although a great increase in transmission capacity is forecast with the new techniques, a much broader scope of coordination will be necessary to control interference.**

INTRODUCTION

MILITARY requirements for global communication trunks are continually increasing because of the demands for higher capacity, improved reliability, and greater flexibility imposed by the addition of new weapons systems. The advent of the satellite relay in orbit is agreed by all to be the logical medium to supplant the over-burdened, erratic, and vulnerable ionosphere for long-distance communications. The possibility of line-of-sight relay transmission over thousands of miles offers promise of large transmission bandwidths to be exploited, freedom from violent fading and multipath distortion, and protection from interference by virtue of narrow antenna beamwidths at microwave frequencies.

From past experience, it is known that once a new propagation medium is proven useful, little time passes before the medium is supporting a large volume of traffic. In view of a number of factors associated with satellite communications, it is likely that mutual interference problems may occur early in the history of its development. Some of these factors are listed.

1) The location of trunk terminals for military applications must be well distributed in geography and reasonably mobile, in order to meet the changing traffic demands of the geopolitical and military situation.
2) The frequency spectrum considered most suitable for this type of application (about 2000 to 10,000 mc) is already in general usage by ground and airborne radar systems and conventional radio relay.
3) Communication, tracking, and telemetering systems for space vehicles are likely to occupy parts of this spectrum.
4) The large transmission distances involved between relay and ground stations may require ground transmitter power, antenna sizes, and receiver sensitivities close to the limit of the art. This can cause local interference problems in the vicinity of the ground stations.

It is not the purpose of this paper to suggest that rigorous controls be imposed on the early experimental systems planned for space relay applications. Actually a great deal remains to be learned about satellite technology and space electronics before communication planners will be able to select and exploit the system best suited for their requirements. On the other hand, the potential use of this medium is sufficiently great to warrant some early consideration of the basic factors affecting efficient spectrum utilization and interference. The remainder of this paper is devoted to a preliminary examination of these factors and some predictions about the type of coordination which may ultimately be required when space communications come into being on a broad scale.

TYPES OF GENERAL COVERAGE RELAY SYSTEMS

A number of different satellite relay systems have been proposed for reliable, high-capacity worldwide communications. A brief review of their distinguishing features will be presented. Basically, a number of relays are placed in specified orbits in such a way that a large fraction of the earth's surface is visible to at least one relay, with sufficient coverage overlap to make communication possible between any two points (via suitably located ground relay stations, if multiple hops are required). These relays may be found in a variety of orbits, including equatorial, polar, and inclined. Usually circular orbits are contemplated, but highly eccentric orbits may be utilized for special purposes.

One particular equatorial orbit which has gained con-

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† Ramo-Wooldridge, a division of Thompson Ramo Wooldridge, Inc., Los Angeles, Calif.

siderable interest is the so-called 24-hour orbit.[1] A satellite in this orbit has a 24-hour period, so that it appears fixed with respect to a point on the earth's surface. For simplicity, a relay in a 24-hour orbit is often called a 24-hour relay (or satellite).

The relay basically can be either of two types, active or passive. The active relay contains a receiver and transmitter, and reradiates the received signal at a higher power level, either directly or after demodulation and remodulation. The passive relay simply reflects or scatters a portion of the incident energy and, accordingly, does not change the modulation in any way (except for possible distortion effects). Some examples of passive reflectors which have been proposed for use include: 1) large sphere, 2) small spheres or ionized particles, 3) flat plate, and 4) chaff in the form of half-wave dipoles. The large sphere has the advantage of being nondirectional, and may be taken as the standard for comparing the scattering cross section of other reflectors. The half-wave dipole is useful because it achieves a large gain in scattering cross section per unit mass, due to resonance effects. In addition, the scattering from a large number of randomly oriented dipoles is essentially nondirectional. The flat plate, when large, also achieves a large gain in cross section but is extremely directional, a severe limitation for use in a system for worldwide coverage. A summary of the scattering cross section per unit mass for several types of passive reflectors is given in Table I.

TABLE I

SCATTER CROSS SECTION FOR A GIVEN MASS

Type	σ Square m	Relative gain over large spherical shell—db
Large spherical shell 17-meter radius	925	0
100 spherical shells 1.7-meter radius	925	0
2.5×10^{6} spherical shells 1.1-cm radius	925	0
3×10^{8} spherical shells 10^{-1}-cm radius	6.6×10^{-2}	−41
7×10^{13} solid shot 10^{-3}-cm diameter	2.4×10^{-10}	−126
9.4×10^{9} half wave dipoles 10^{-3}-cm diameter	1.9×10^{7}	+43
Flat circular plate 34-meter radius	8.5×10^{9}	+70

Mass = 100 kg of Aluminum
Thickness = 10^{-3} cm
Wavelength = 0.1 meter

In at least one proposed system,[2] a large number of passive relays in orbit is found necessary to ensure complete coverage at all times. This leads to another way of classifying relay systems, *i.e.*, as to whether the relays are "discrete" or "distributed." The former achieves full coverage by optimum positioning of relatively few relays in orbit; the latter by placing a large number of simple small relays in orbit essentially at random. In view of complexity and reliability considerations, a distributed system may be expected to be composed of passive relays.

APPLICATIONS OF ACTIVE AND PASSIVE RELAYS

Since the state of the art indicates the feasibility of both active and passive relays, some idea of the applications for which one is more suited than the other will aid in assessing potential mutual interference problems. This requires a meaningful method for evaluating comparative performance capabilities.

A completely adequate method does not appear to exist, the attempt at evaluation being, in a sense, comparable to a choice between apples and pears. Two methods are suggested here, both based on the requirement to produce equal power densities at the ground receiver. The active relay is characterized for the comparison either by its power gain or by its maximum power output, while the passive relay is described by its scattering cross section. These parameters are sufficient to define the electrical performance of the relays.

The notation employed in the following discussion is:

G_T = gain of ground transmitting antenna,
G_R = gain of ground receiving antenna,
g_s = relay antenna gain, assumed equal for transmitting and receiving (active),
K = relay power gain (active),
σ = scattering cross section illuminated by both ground antennas (passive),
λ = wavelength,
R_1 = distance from ground transmitting station to relay,
R_2 = distance from relay to receiving station,
P_R = received power on the ground,
P_T = transmitted power,
P_s = relay power output.

The received power on the ground for the two kinds of relays may be expressed as

$$P_R = P_T \frac{G_T G_R \lambda^2}{(4\pi)^3 R_1^2 R_2^2} K g_s^2 \frac{\lambda^2}{4\pi} \quad \text{(active)}, \tag{1a}$$

or

$$P_R = P_s g_s G_R \lambda^2/(4\pi)^2 R_2^2, \tag{1b}$$

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[1] M. Handelsman, "Performance equations for a stationary passive satellite relay (22,000-mile altitude) for communication," IRE TRANS. ON COMMUNICATIONS SYSTEMS, vol. CS-7, pp. 31–37; May, 1959.

[2] J. R. Pierce and R. Kompfner, "Transoceanic communication by means of satellites," PROC. IRE, vol. 47, pp. 372–380; March, 1959.

depending on choice of characterizing parameter, and

$$P_R = P_T \frac{G_T G_R \lambda^2}{(4\pi)^3 R_1{}^2 R_2{}^2} \sigma \quad \text{(passive)}. \tag{2}$$

Comparison of (1) and (2) shows that passive and active relays for the same application may be compared as described above by one of the two equations:

$$\sigma = K g_s{}^2 \lambda^2 / 4\pi, \tag{3a}$$

or

$$\frac{P_T G_T \sigma}{4\pi R_1{}^2} = P_s g_s. \tag{3b}$$

This comparison is valid provided that noise in the relay can be neglected; otherwise, the active relay must have a greater ouptut power than indicated by the equations.

To illustrate the approximate "break-even" point between active and passive relays, a numerical example will be presented. Suppose that the relay is 7000 km from the ground terminals, which utilize 35-foot diameter dishes at 3 kmc (48-db gain). The ground transmitter power is taken as 10 kw. The relay, if active, has essentially an omni-directional antenna; if passive, is composed of half-wave dipoles. Then (3b) yields

$$10^{-6}\sigma = P_s.$$

Table I shows that the scattering cross section of randomly oriented half-wave dipoles is about 2×10^5 meter^2/kg, for $\lambda = 0.1$ meter. Thus, the active relay will be equivalent to the passive relay, under the conditions assumed, if it has a power-to-weight ratio of

$$P_s = 0.2 \text{ watt/kg}.$$

An examination of possible active relay configurations shows that this value of transmitter power per unit weight appears reasonably typical for systems now under development.

A general comparison of (3a) and (3b) leads to the qualitative conclusions that active relays are to be preferred over passive relays primarily for satellites at great distances, or when high-power ground transmitters and large antenna structures are not available. As a result, without a consideration of mutual interference problems, active relay system design will tend to emphasize low ground power, high relay receiver sensitivity, and a high relay power output. Passive relay systems, on the other hand, may be expected to utilize high ground transmitter power and high ground receiver sensitivity, to overcome limits on the total scattering cross section which can be placed in orbit.

POTENTIAL INTERFERENCE RANGE

The interference problem at the microwave frequencies useful for satellite communication can be severe because of the wide coverage obtained on the ground from high altitudes. Assuming straight-line propagation to the horizon, a satellite will be visible from any point within a circle of radius (measured as degrees of arc along a great circle)

$$r = \cos^{-1} \frac{a}{a+h}, \tag{4}$$

where a = radius of earth and h = altitude of orbit. Thus, any microwave source within this circle is a potential interference source. As examples, satellites at altitudes of 500 miles and 22,000 miles (24-hour orbit), would be visible within circles of radius 27° and 81°, respectively.[3] For an equatorial orbit, r would correspond to the limiting latitude in the northern and southern hemisphere.

In the case of the 24-hour relay, interference reduction by a highly directional antenna in an active satellite is feasible, but requires precise attitude stabilization to maintain contact with a selected ground station. For application in a worldwide communications system, however, high-gain antennas cannot be used in the relay without seriously restricting the service coverage per satellite. Since the earth subtends an angle of about 18° at the altitude of a 24-hour satellite, an antenna gain of about 20 db is the maximum which can be utilized with no restriction in coverage.

SYSTEM CAPACITY WITH HIGH TRAFFIC DENSITY

In contrast to conventional radio relay systems, the relative cost of the relay "installation" will be large, since it involves all the problems of placing satellites in orbit. For maximum benefit to be derived from the investment, it is logical to expect that the relay will be required to serve several terminals simultaneously. If this occurs without global coordination, and if a sufficient number of independent sources exist, the resulting interference problem will be similar to the case of relaying a signal in the presence of excessive random noise. Although this assumption may be pessimistic, it is a valid limitation to discuss, especially since the satellite relay transmission parameters are not easily changed after the orbit is established.

Under these conditions it will be possible to trade information rate in each individual ground terminal to reduce the effect of the interfering signals. Examples of such trades are well known as in the case of PCM, frequency-modulation, etc. Modulation systems of this type exhibit a threshold characteristic which depends on a minimum permissible value of signal-to-interference ratio for proper performance. In principle, a more general limit of trading can be derived from the Shannon formula for information rate,

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[3] The often-used 4/3 correction factor on the earth's radius of 4000 miles has not been introduced here.

$$R = B \log \left(1 + \frac{S}{N}\right) \text{ bits/second,} \tag{5}$$

where B is the signaling bandwidth, S is the average power and N is the total interference power. Under conditions of heavy interference, (5) may be approximated by

$$R = 1.44B \left(\frac{S}{N}\right) \text{ bits/second,} \tag{6}$$

thereby indicating that a linear trade of bandwidth for interference reduction is the best that can be achieved. As an example, consider the capacity available to an individual communication system derivable from a single relay serving a number of other systems. Let C be the capacity, expressed as the number of permissible voice channels with a tolerable SNR of S/N_v per channel, let B denote the total system bandwidth, and B_v the voice channel bandwidth. Then (6) is stated for information rate with and without interference,

$$CB_v \left(\frac{S}{N_v}\right) = B \frac{S}{N} \cdot \tag{7}$$

Typical values for the parameters are taken as

$$B = 3000 \text{ mc},$$
$$B_v = 3 \text{ kc},$$
$$S/N_v = 100 \text{ (20 db)}.$$

Table II indicates the resulting number of satisfactory voice channels for various levels of interference. It is obvious from the table that a single relay can be expected to provide a reasonably large capacity, on any given circuit, only if potential sources of interference are properly coordinated so that heavy interference cannot occur. In fact, under severe interference conditions (signal-to-interference ratio less than −10 db) a single TV channel (1000 voice channels) cannot be supported on one circuit.

TABLE II

Capacity per Circuit of a Single Satellite Relay

Number of voice channels	Signal-to-interference ratio
1000	−10 db
100	−20 db
10	−30 db

In practice, the matter is further complicated because the active relay cannot exploit the trade to the extent indicated. Unless demodulation, decoding, and remodulation are employed in the relay (with consequent problems of complexity and reliability) it must exhibit linear transmission properties up to the highest level of interference to be encountered. Thus, for any practical design a limit of tolerable interference level at the relay is inherent, and may be viewed as another type of system threshold. Passive relays, on the other hand, are not subject to such saturation effects, and would permit the types of trade indicated above by manipulation of the ground terminal equipments.

When full coordination of the various sources of radiation is achieved, the single satellite can be fully loaded over its entire available frequency band (within transmitter power limits). The total bandwidth is thus subdivided into noninterfering voice channels, the available number being

$$\text{number of voice channels} = \frac{B}{B_v} = 10^6. \tag{8}$$

Comparison with Table II shows the many orders of magnitude improvement available with the aid of full coordination, a conclusion which applies equally to active relays and discrete passive relays.

Interference from Other High-Power Microwave Sources

Because a satellite relay is visible over a large portion of the earth's surface, as discussed above, interference from high-power microwave transmitters, radar for example, is potentially serious. To indicate the situation that might be encountered, a particular satellite communication system, namely an active relay in a 24-hour orbit, will be examined. An advantage sometimes supposed for the active relay is the relatively low ground power required. This advantage will be examined quantitatively by choice of typical system parameters. Suppose that:

relay antenna gain = 20 db,
ground antenna gain = 40 db,
carrier frequency = 3000 mc,
information bandwidth = 3 mc = 1000 voice channels,
receiver noise figure = 0 db, and
desired SNR = 23 db.

A simple calculation shows that a transmitter power of about 60 watts will suffice when the only limitation on receiver performance is thermal noise.

To assess the potential interference from radar transmitters, a reasonably typical radar set will be presumed with the parameters:

average power = 1 kw,
average antenna gain = 0 db,
emission bandwidth = 1 mc, and
carrier frequency = any point within S band (assumed nominally 2000–5000 mc).

It will be assumed that the various radars visible to the relay are uniformly spread over the 3000-mc band.

The communication system will be considered marginal when the total interference is 20 db below the signal. If n = total number of visible active radars, marginal operation occurs when

$$n \frac{1000 \text{ watts}}{3000 \text{ mc}/3 \text{ mc}} = \frac{60 \text{ watts}}{100} \cdot 10^4,$$

or

$$n = 6000.^4$$

Since practically a full hemisphere of the earth's surface is visible to the 24-hour relay, corresponding to an average area of perhaps 25×10^6 square miles, an undesirable level of interference may be expected if the density of active radars exceeds an average of one every 4000 square miles. Correspondingly higher densities must exist to cause excessive interference as the signal level is increased over the minimum requirement.

The radiation received at the relay from interfering sources is much less of a problem with passive relays, because of the higher ground power and antenna gain required to overcome thermal noise. For example, Pierce[2] discusses use of a 100-foot diameter sphere as a passive relay in orbit at 3000 miles altitude. Assuming 70-db antennas at 10 kmc (130-foot diameter), a low-noise receiver (noise temperature = 20°K), and a mean path distance from relay to terminal of 5000 miles, a transmitted power of about 20 kw is necessary to produce a 20-db SNR in a 3-mc bandwidth. Compared with the minimum required power calculated above for the active relay, the effective radiation is about 55 db higher in intensity.

Although interference at the relay does not appear a significant problem with passive satellites, interference radiated directly into the ground receiver may be intolerable. For example, the antenna for the ground receiver in the passive relay system discussed above may have a sidelobe 50 db below the main beam, yielding an effective aperture of only about 10^{-2} square meters. Nevertheless, an interference power comparable to the thermal noise power is intercepted by the sidelobe of the antenna, if the incident power density is about 10^{-13} watts/meter^2. This corresponds to the intensity produced at a 5-mile distance by 0.1 mw radiated isotropically over a line-of-sight path.

It should be noted that the expected condition of radars distributed uniformly over the total available frequency band precludes useful application of the interference reduction techniques previously discussed, since the interference power is directly proportional to the signalling bandwidth used. This offsets exactly the linear trade normally obtained with such techniques. Thus, it may be concluded that complete coordination of the radar and satellite communication frequencies is essential to take advantage of the relatively low power found sufficient for an active relay system, when interference is not present. In addition, coordination may be the only method of preventing overloading when extremely low-noise ground receivers are employed for passive relay systems.

CONCLUSION

The discussion has indicated that simultaneous use of active and passive relay systems with their differing power requirements, and without proper coordination between the two systems, can lead to severe mutual interference. That is, high-power transmitters employed in one system can cause undesirable levels of interference in the receivers of the other system.

When heavy interference is likely with either an active or a discrete passive relay, the capacity of the relay system is severely limited by a requirement for reliable transmission. From a practical point of view, therefore, the ultimate usefulness of active relays for worldwide communication is questionable. Because of the capacity limitation due to interference inherent with discrete relays, the authors foresee a trend toward development of "distributed" passive relay systems. Such systems can provide a large capacity, by virtue of a large number of available reflectors in orbit, with the opportunity for assigning channels on the basis of location of the relay in space as well as radio frequencies.

The calculations indicating possible intolerable interference from radar and other microwave sources further substantiates the probability of a trend away from active relays. Since a low ground-power level does not appear feasible due to the interference level resulting when radars are not under control of a supervising agency, the main advantage of active systems is thereby negated. The great advantage of passive systems, *i.e.*, simplicity of the device in orbit, will, in the authors' opinion, win out for military communication systems, in which reliability is foremost. Thus, the final result of programs to develop a reliable long-range satellite communication system may well be a passive medium having the desirable features of the ionosphere, and few of the disadvantages.

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[4] The presence of harmonic and spurious emissions from the radar sets can effect a reduction of this number. A. H. Ryan, "Control of microwave interference," IRE TRANS. ON RADIO FREQUENCY INTERFERENCE, vol. RFI-1, pp. 1–10; May, 1959.

Passive Satellite Communication*

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Summary—**A number of communications capabilities are noted as desirable goals and the characteristics of passive satellite reflectors are cited as a means of achieving them. Goals noted are the attainment of transoceanic ranges, high communications traffic per satellite, low launching and operating costs, high reliability, simplicity of operation, and long useful life of satellites.**

The properties of various types of passive reflectors are discussed and the characteristics of the sphere are shown to be particularly advantageous. Passive satellite communications systems are compared with current tropospheric scatter systems and are shown to be one order of magnitude better, using power per channel per unit range as a criterion.

Some attention is given to Doppler shift as it bears upon the problem of transmitter-receiver channel coordination, satellite acquisition, and tracking.

I. INTRODUCTION

SINCE the advent of the first Russian Sputnik much has been written, proposed and accomplished in the field of earth satellites. The use of satellites as repeaters for long-range communications systems has been proposed and numerous papers have been presented based upon the use of active satellite repeaters. The emphasis on, and interest in, passive repeaters, however, has lagged and the current literature has largely dealt with the global coverage of various orbit configurations.

The purpose of this paper is not to propose additional system configurations but to examine the feasibility of passive satellite relays in general and to discuss to some extent the approaches available for their realization.

The material will be presented in terms of the characteristics of a single ground point-to-point circuit during the transit of a single relay satellite. The systems problems involving a multiplicity of satellites has been very well treated by Pierce and Kompfner.[1]

The approaches described are directed toward the attainment of the following goals:

1) Attainment of transoceanic ranges of the order of 6000 nm maximum.
2) Availability of the relay to a large number of users.
3) Attainment of a feasible satellite launching cost per channel.
4) Attainment of a feasible operating cost per channel per mile of range.
5) Attainment of a long satellite orbit life.
6) Development of a rapid and simple satellite acquisition and tracking system.
7) Attainment of high satellite reliability by elimination of active mechanical and electronic devices in the satellite.

II. ASSUMPTIONS

All numerical values presented were based on first order feasibility-type calculations. Since no equipment details such as line losses, feed polarizations, etc., were considered, comparable propagation refinements such as radio horizon corrections, Faraday rotation, change of cosmic noise levels with zenith angles, etc., were not included.

The combined effective noise temperature for receiving systems directed from earth to space was assumed to be 10°K,[1] system directed from satellite to earth, 300°K. Signal to noise was always assumed 20 db.

Due to aperture-to-medium coupling losses the maximum gains assumed possible for ground-based antennas was a conservative 50 db.[2] Satellite-borne antennas were assumed to be omnidirectional unless otherwise noted.

The maximum available satellite-borne primary power was assumed to be a maximum of 1 kw for the immediate future. In addition, the maximum radiated power for satellites was assumed to be 100 watts, an efficiency of 10 per cent over all.

A desirable operating spectrum for earth-space systems was assumed to be from 2 to 10 kmc. The lower limit was selected on the basis of antenna aperture and gain considerations as well as ionospheric effects; the upper limit is based on atmospheric abosrption. Discussions of traffic capacities are in terms of channels, a channel being arbitrarily defined as 4 kc. It is further assumed that the circuitry and primary power related to a channel of transmission capacity in a satellite require an average of 5 lbs of total payload. For convenience the carrier frequency for all systems was assumed to be 3 kmc, or $\lambda = 0.1$ meter.

The system configuration assumed for power calculations will consist of the extreme condition of the satellite midway between two ground stations separated by a great circle distance of 6000 nm. This is roughly one fourth the circumference of the earth. If the satellite is to be visible from both stations it must be at an altitude of at least 1600 nm.

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[1] J. R. Pierce and R. Kompfner, "Transoceanic communications by means of satellites," PROC. IRE, vol. 47, pp. 372–380; March, 1959.

[2] Lichman, "Passive Relay Communications," Hughes Aircraft Co., Culver City, Calif., Rept. No. 4551/2; January 9, 1959.

On the average it is expected that the satellite will be in view 10 or 15 degrees either side of the mid-position since ground stations need not be at the extreme range. The slant range from either ground station to the satellite will be 4000 nm. This will be assumed to be the geometric mean range for all calculations. The antenna gain of each ground station will be assumed identical.

Doppler calculations will be based on the average great circle range between stations of 3300 nm. This will provide a maximum mutual visibility of ±15° either side of the mid-point along the great circle arc joining the stations at an altitude of 1600 nm.

III. Passive Relays

All passive satellite relay systems may be characterized by the radar range equation. The energy reflected by the satellite must be received by the distant receiver which, in the communication case, is not situated at the transmitter site. Discounting the possibility of using optical frequencies or higher due to atmospheric absorption effects, ranges will be sufficiently high to place the satellite in the far fields of both transmitting and receiving antennas. Consequently, simplified formulas available in the literature may be used to represent the radar cross section of the satellite.

In addition to the far field criterion, further simplifying assumptions must be made with regard to the satellite dimensions to avoid complexities due to diffraction effects which occur when one or more dimensions of the reflector are less than one wavelength. In general, for this condition, the cross section reduces rapidly as the dimensions become smaller relative to a wavelength although harmonic peaking usually occurs.

The radar range equation for the CW case may be written:

$$P_T = \frac{(S/N)KT_eB(4\pi)^3R^4}{G^2\sigma\lambda^2} \tag{1}$$

where

P_T = average transmitter power, watts
G = antenna gains
R = geometric mean range, nm
T_e = effective noise temperature °K
B = bandwidth, cps
K = Boltzmann's constant
λ = carrier wavelength, nm
σ = satellite cross section, square nm.

In general the cross section, σ, may be viewed as an effective receiving cross section, A_r, multiplied by a retransmission gain G_T:

$$\sigma = A_rG_T. \tag{2}$$

In the cases under consideration here, where dimensions are greater than a wavelength, A_r may be approximated by the physical area projected normal to an incident plane wave. One exception is the very thin half-wave dipole which will have a physical area much less than its electrical capture area.

The cross section of various simple geometrical figures are noted in Table I.[3] This list is far from exhaustive. Many other reflectors may be conceived of, such as dipole arrays which redirect energy to the source (Van Atta array), dipole arrays with internal switching to redirect energy in a desired direction, Luneburg lenses, inverse Luneburg lenses, helices, etc.

TABLE I

	Cross Section	Retransmission Gain	Directive Property
Reflector	σ	G_T	
Half-wave dipole	$\frac{2.9\lambda^2}{4\pi}$	1.7	energy directed normal to long axis
Sphere	A	1	isotropic*
Cylinder	$\frac{\pi LA}{\lambda}$	$\frac{\pi L}{\lambda}$	energy tends to be directed normal to long axis
Flat plate	$\frac{4\pi A^2}{\lambda}$	$\frac{4\pi A}{\lambda^2}$	angle of reflection equals angle of incidence
Parabolic dish, dipole array (broadside)	$\frac{4\pi A^2}{\lambda^2}$	$\frac{4\pi A}{\lambda^2}$	must be directed within one beamwidth of transmitter, return to transmitter
Corner reflector	$\frac{7\pi A^2}{\lambda^2}$	$\frac{7\pi A}{\lambda^2}$	return energy to transmitter

* As an opaque sphere constitutes an obstacle, gain will occur directed back of the sphere along a radial line to the source. Similar effects will occur with irregular obstacles.

Since one of the desired goals noted in the Introduction was the attainment of maximum satellite reliability through the elimination of active devices in the satellites, many of the items noted in Table I may be disqualified.

All of the reflectors have some directional properties, this, of course, being related to the retransmission gain. In most cases, if advantage is to be taken of the gain, some means must be provided, either mechanical or electrical, to redirect the energy to the desired receiver. If these items are discarded, the only remaining devices would be the sphere and corner reflector in Table I and the Van Atta array[4] and inverse Luneburg lens[5] noted in the text.

The inverse[6] Luneburg lens is conceived of as a variable dielectric material which rises to infinity at the center, the incident radiation being returned to the direction from which it was received. Assuming this device is realizable, it is probable that its inherent weight would disqualify it as compared to a spherical nest of corner reflectors. The nested corners could be composed of a lightweight, inflatable metalized plastic.

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[3] "Reference Data for Radio Engineers," International Telephone and Telegraph Corp., p. 804; 1958.

[4] E. D. Sharp, "Properties of the Van Atta Reflector Array," Rome Air Dev. Center, Rome, N. Y., RADC TR-58-53; March, 1958.

[5] R. E. Webster, "Radiation patterns of a spherical Luneburg lens with simple feeds," IRE Trans. on Antennas and Propagation, vol. AP-6, pp. 301–302; July, 1958.

[6] Emerson and Cummings, Inc., Canton, Mass., personal correspondence.

The Van Atta array could also be realized in a lightweight configuration. The dipole elements and interconnections could be plated on a lightweight inflatable balloon structure.

For the system geometry selected it is doubtful that it would be worthwhile to develop a reflector system having gain, yet not require active elements for orientation. In order for the 3-db points of the main return beam to subtend a 6000-mile range, the aperture of the reflector would have to be selected to provide a beamwidth of 2π steradians, a gain of 3 db. Using the corner reflector gain equation of Table I at 3 kmc ($\lambda = 0.1$ m), A becomes 10^{-3} square meters. This results in a cross section of 2×10^{-3} square meters.

On the other hand the sphere cross section is substantially independent of wavelength provided its radius exceeds the length of one wavelength. As it is a mechanically symmetrical object in which only the surface is of interest, there is no reason to believe that human ingenuity cannot expand its dimensions to provide extremely high cross sections.

If the 6000-mile great-circle range is to be achieved by the use of a corner-reflector type satellite in synchronous orbit at an altitude of 22,000 nm, the 3-db points of its retransmission beam must subtend 11,200 miles. Its beamwidth is then $\frac{3}{4}$ steradian corresponding to a gain of 16 or 12 db. Using the reasoning employed previously its cross section would be 0.128 square meter. An equivalent sphere would only require 12 db more of cross section or 2 square meters. A much larger sphere could be readily placed in orbit.

The only item in Table I of dimension less than a wavelength is the half-wave dipole. This has the advantage of having a cross section greater than its physical area but has directivity disadvantages as well as a small individual cross section. The treatment of randomly oriented quantities of dipoles is beyond the scope of this paper but is treated in the literature.[7]

IV. PASSIVE VS ACTIVE

No doubt the reader is concerned with the fact that passive satellite reflectors are even being considered. It is obvious to everyone familiar with transmission systems that only minute quantities of power are required for active systems as compared to passive. This question will be examined a little more critically here.

The active and passive satellite relay system for the specific configuration assumed here may be readily compared by the use of the range equations for the two systems. The range equation for the passive system is already stated in (1), which gives the ground transmitter radiated power requirement. The range equation giving the ground radiated power requirement for the active case need only consider the one-way requirement from ground to satellite. This is

$$P_T = \frac{(S/N)KT_eB(4\pi R)^2}{G\lambda^2}. \qquad (3)$$

Eq. (3) has been written subject to the previously stated assumptions of unity gain for the satellite antenna and the system configuration described. The symbols have the same meaning and units as noted in connection with (1).

The technique of comparison will be to assume that the design parameters of both systems are identical, including the ground radiated power. The effective receiver noise temperatures must be different in accordance with the assumptions. Eq. (3) will then be divided by (1) and the appropriate cancellation performed. The resulting relationships will then specify the cross section of the passive satellite required to match the performance of the equivalent active system.

$$\sigma = 4\pi\left(\frac{T_{ep}}{T_{ea}}\right)\frac{R^2}{G}. \qquad (4)$$

The subscripts p and a refer to the passive and active systems respectively. Substitution of the assumed parameters yields a required cross section of approximately 60 square miles.

It is doubtful if such a reflector cross section will be realizable within the next decade or more. At present lightweight inflatable balloons of 100 feet in diameter have been fabricated and will be launched into 1000-mile-high orbits within the next year.[8] It is probable that 1000-foot-diameter balloons will eventually be placed in orbit. The cross section of such a reflector will be 2.1×10^{-2} miles. A foreseeable passive system requiring no active equipment in the satellite will then require 34 db of system gain over the active systems in order to match the capability of active systems. A passive system using 100-foot-diameter spheres will require 54 db of additional system gain.

At first glance the above figures appear to reinforce the intuitive judgement of the communication engineer that passive systems are not feasible. However, before discarding them on this basis one should inquire as to the current value placed upon the attainment of long-range communications.

There are many areas of the world where it is not feasible, desirable, or even possible to reliably bridge long ranges by wire, transoceanic cable, a multiplicity of microwave links or HF. In order to meet this need, tropospheric and ionospheric forward-scatter systems have been developed and placed in operation at great expense.

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[7] Van Vleck, Bloch, and Hamermesh, "Theory of radar reflections from wires or other metallic strips," *J. Appl. Phys.*; March, 1947.

[8] "Next decade in space," *Aviation Week*; June 22, 1959.

The premium placed upon long range communications in terms of required radiated power per channel per mile of range may be readily calculated, for example, for a tropospheric scatter system having a range of 400 nm. Current literature[9] indicates that such a system will have a loss of 98 db in excess of free space, and will require 3.3 kw per channel of 4-kc bandwidth.

Eqs. (1) and (3) may be used to calculate power required per channel per mile of range. For satellite systems these calculations are summarized in Table II.

TABLE II

System Type	Radiated Power/Channel	Power per Channel Mile
Active satellite, satellite-to-ground circuit	0.43 mw	0.07 μw
Active satellite, ground-to-satellite circuit	13 mw	2.5 μw
Passive satellite, 100-foot-diameter sphere	3.9 kw	0.6 w
Passive satellite, 1000-foot-diameter sphere	39 w	5 mw
Tropospheric scatter, 400-mile range	3.3 kw	8.2 w

Table II shows that, although the 100-foot satellite is not competitive with an active system, it is certainly competitive with a tropospheric scatter system. The 15 scatter links required for a 6000-mile range would require a total expenditure of 495 kw as compared to 3.9 kw for the passive system.

In terms of availability of power sources and the value attached to long-range communications systems as evidenced by the development of scatter systems, it is apparent that a passive system based upon the use of 100-foot-diameter spheres is feasible.

V. Traffic Capacities

A desirable goal is the attainment of a maximum traffic capacity through the realization of a large number of individual channels per satellite relay. This enhances its value in terms of the number of users to which it may be made available.

It was pointed out under the assumptions that a spectral space extending from 2 to 8 kmc should be available based on propagation and antenna size considerations.

The passive spherical satellite will be capable of handling channels allocated throughout this space, as its cross section is essentially not dependent on frequency. As noted, its cross section is frequency dependent when wavelengths become comparable to diameter. In the case of the 100-foot-diameter sphere this would reduce the lower frequency end into the HF region of the spectrum. The upper-frequency limit would probably be based upon the surface irregularities due to the practical fabrication difficulties. The upper propagation band of 10 kmc would require that these irregularities be considerably less than 3 cm to avoid phase cancellation and multipath.

If 50 per cent of the 8-kmc spectral space available is used for 4-kc channels, the traffic capacity of the passive spherical satellite system would be one million channels.

Available data on the 100-foot-diameter sphere being developed by NASA indicates that its weight will be approximately 100 pounds.[10] As surface area of the sphere having constant weight per unit surface rises with the square of diameter, a 1000-foot-diameter balloon should weigh five tons. It is probable that payloads of this magnitude will be capable of being placed in orbit within the next few years.

If five tons is considered to be a reasonable payload within the foreseeable future, this number may be used in estimating the feasible number of channels of an active satellite repeater. For this purpose it will be assumed that total antenna primary power casing, frequency standards, microwave oscillators and associated circuitry required per channel in an active satellite relay will result in an average weight of five pounds per channel. On this basis the traffic capacity per satellite would be 2000 channels.

In terms of the passive satellite of 1000-foot-diameter the average weight per channel would be 0.16 ounce or 4.5 grams, less than the weight of a one-karat jewel. It is difficult to conceive of an active system of comparable channel capacity being developed in the foreseeable future.

VI. Longevity

With the exception of the solid dielectric type reflectors, the passive satellite reflectors considered are concerned only with electromagnetic reflecting surfaces of the order of 1000 angstroms thick. One intuitively feels that drag effects, even in the tenuous atmosphere of space, should disturb the orbits of such objects. Quantitative analyses of the orbit of low-density satellites must consider the particle density estimates derived from sounding rockets, the aerodynamics of objects in these low-density atmospheres and the orbit perturbations determined by the numerical integration with time of the resulting drag forces. Since orbit parameters in the simple two-body case are determined by the total particle energy per unit mass, a useful parameter in such calculations is the ratio of the projected area, normal to velocity, to the mass of the satellite.

By ordinary drag theory the loss of kinetic energy per unit mass for each orbit revolution is the product of a constant and the ratio of area to mass. This loss of kinetic energy has the effect of making the total energy more negative, which is the equivalent of reducing the radius of a circular orbit. Every reduction of energy increases the drag forces at a lower altitude and the vehicle gradually spirals into the dense atmosphere and burns up.

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[9] "Reference Data for Radio Engineers," International Telephone and Telegraph Corp., p. 761; 1957.

[10] NASA data; June, 1959.

Calculations based on best estimates of high-altitude atmospheric densities indicate the life[11] of a 100-pound, 100-foot-diameter balloon to be 25 years for an orbit having an apogee of 1600 miles and a perigee of 1000 miles altitude. A similar solid dielectric sphere having roughly the density of water would have a weight of about 300 tons and consequently have a much greater life. On the other hand should the balloon orbit perigee be only 500 miles, its life expectancy would be only 2 months.

By contrast, using a linear approximation for the solid dielectric sphere, its life for a 500-mile perigee would be 1000 years.

In the case of the solid satellite, experience would indicate that micrometeoric erosion does not constitute a problem (with the possible exception of its effect on solar power cells). On the other hand, one would expect a considerable effect in the case of the balloon-type reflector composed of rather diaphanous structural material supporting a thin metallic reflecting film. No doubt considerable experience will have to be gained through the actual launching of balloons into orbit before a suitably rigidized lightweight balloon can be launched with a realistic life expectancy of several years.[12]

The above considerations of life-limiting factors do not reveal any fundamental limitations. In the final analysis a spherical reflecting satellite of arbitrarily low ratio of area to mass, fabricated of stainless steel, if necessary, may be placed in a virtually permanent orbit. The fundamental limitation in this case is, of course, the state of the development of rocket technology and the availability of the economic resources necessary to launch massive bodies into orbit.

VII. Acquisition and Tracking

There are various approaches to this problem depending upon the orbit system configuration adopted. In this case it is assumed that the ground stations are separated by 3300 nm. The detailed geometry of the configuration is described in the Appendix.

It will be assumed that the satellite schedules of arrival over the horizon are known within $\pm 2.5°$ of the calculated position and within a few seconds in time. The assumed antenna gain of 50 db at 3 kmc provides a beamwidth of $\frac{1}{2}°$. A channel operating at 300 mc will provide a gain of 30 db with the same antenna at a beamwidth of 5°.

It is demonstrated in the Appendix that the Doppler frequency changes approximately linearly from acquisition to release. The radius of the circle of mutual visibility is about 1200 miles. The slant range to the satellite will be about 3600 miles, which at a 5° beamwidth subtends about one fourth of the zone through which the satellite will pass. As the upper Doppler frequency shift is 93 kc, this shift should change over the first quarter of the shift down, or 23 kc.

In order to assure Doppler acquisition, the 4-kc tracking bandwidth could be located in discrete bands throughout the 23-kc band in order to assure frequency acquisition. It is shown in the Appendix that the minimum bandwidth required for acquisition is 53 cps. Use of four 100-cps bandwidth filters should allow ample acquisition bandwidth. At a 10-db signal-to-noise ratio the power required would be 3.9 kw.

Having acquired the satellite in both space and frequency the tracking circuit may lock on and follow. Under tracking conditions the Doppler bandwidth may be narrowed considerably for a higher signal-to-noise margin. Standard multiple interferometric or conical scanning techniques may be used. Communications receiving channels may be locked to the tracking frequency so that automatic coordination between transmitting and receiving frequencies is effected. The effect of Doppler shifts on expanding or contracting transmitted bandwidths is shown in the Appendix to be negligible.

VIII. Conclusion

The use of individual passive satellites as reflecting media for transmission of microwave communications over ranges from 3000 to 6000 miles has been examined from the point of view of attaining a series of desirable goals. It is concluded that the lightweight spherical satellite provides communication for a large number of users on a weight per channel basis which cannot be met in the foreseeable future by existing active satellite repeater techniques. In addition, due to the absence of active satellite-borne components, it has a reliability feature which cannot be matched by other types of systems.

The attainment of long satellite orbit life of the order of decades will require extensive development but no fundamental scientific difficulties must be overcome.

The operating cost per channel per mile of range will be initially one order of magnitude better than present state-of-the art long-range tropospheric scatter systems and may be greatly improved through the development of large diameter satellites and ground antennas having gain through the atmosphere in excess of 50 db.

Satellite acquisition tracking and channel frequency coordination may be accomplished by the adaptation of Doppler state-of-the art techniques.

Appendix

Doppler Calculations

System Geometry

Satellites have been assumed to be on the surface of a sphere of altitude Z_0. Transmitter and receiver station

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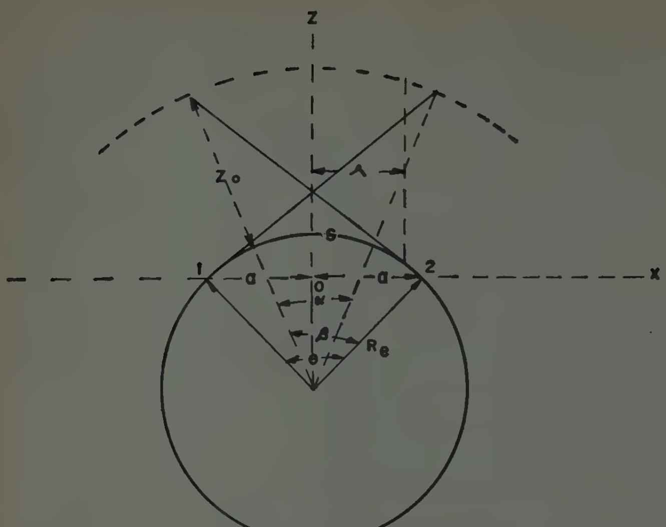
[11] NASA data, June, 1958.

[12] Recent unpublished analyses of the effects of solar radiation pressure on the orbits of low average density satellites indicate a life limitation because of this factor. These calculations show strong dependence of the life on the satellite on altitude and inclination of the orbit plane to that of the earth's equation. However, in the case of 100-foot-diameter balloons having altitude of 1600 nautical miles, inclined as much as 30° to the Equator, solar radiation pressure appears to have only a negligible effect on longevity.



S=great circle distance between stations 1 and 2
θ=angle subtended at center of earth by station 1 and 2
β=horizon range of satellites in angular measure
α=range of satellite mutual visibility in angular measure
r=mean radius of mutual visibility
a=semichord range between station 1 and 2
R_e=radius of earth
Z_0=altitude of satellites

Fig. 1.

are shown in Fig. 1 at points 1 and 2. Given:

$$R_e = 4000 \text{ nm}$$

$$S = 3260 \text{ nm.}$$

Find: r and a;

$$\theta = \frac{S}{R_e} \tag{1}$$

$$a = R_e \sin \theta/2 \tag{2}$$

$$r = \left(R_e + \frac{Z_0}{2}\right)\left(\frac{\alpha}{2}\right) \tag{3}$$

$$\alpha = \theta - 2(\theta - \beta) \tag{4}$$

$$\beta = \cos^{-1} \frac{R_e}{R_e + Z_0}. \tag{5}$$

Then

$$r = 1200 \text{ nm maximum}$$

$$a = 1960 \text{ nm.}$$

Doppler Geometry

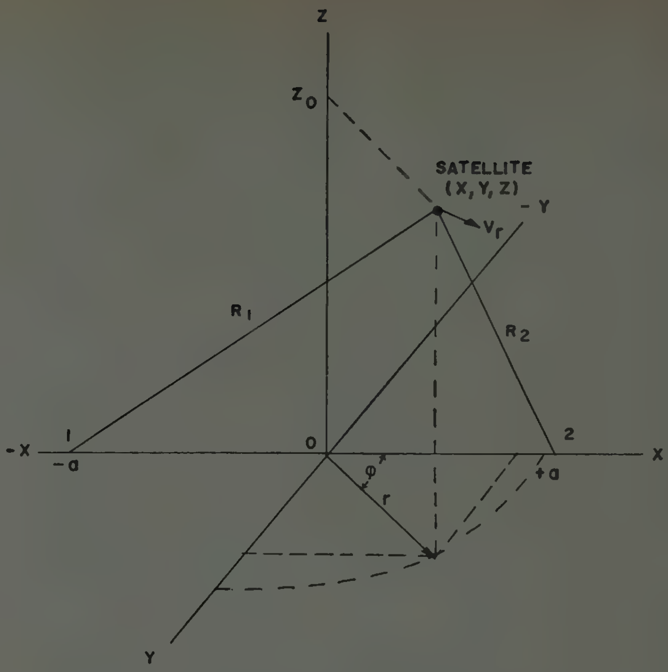
A Cartesian coordinate system is used based upon parameters derived from Fig. 1. This system is shown in Fig. 2.

The radial velocity components may be obtained from the total derivative with respect to time for the radial ranges:

$$R_1^2 = Z_0^2 + y^2 + (x + a)^2 \tag{6}$$

$$R_2^2 = Z_0^2 + y^2 + (x - a)^2 \tag{7}$$

$$R_1 V_{r_1} = yV_y + (x + a)V_x \tag{8}$$



v_r=velocity of satellite approximated as parallel to XY plane
r=displacement of satellite along V_r direction from 0
ϕ=angle of intersection of satellite path to chord between stations 1, 2
R_1=range from station 1 to satellite
R_2=range from station 2 to satellite

Fig. 2.

$$R_2 V_{r_2} = yV_y + (x - a)V_x. \tag{9}$$

Cylindrical coordinates may be derived based upon the transformations:

$$X = r \cos \phi \tag{10}$$

$$V_x = V_r \cos \phi \tag{11}$$

$$V = r \sin \phi \tag{12}$$

$$V_y = V_r \sin \phi. \tag{13}$$

The transformed equations become

$$R_1^2 = Z_0^2 + r^2 + a^2 + 2ar \cos \phi \tag{14}$$

$$R_2^2 = Z_0^2 + r^2 + a^2 - 2ar \cos \phi \tag{15}$$

$$R_1 V_{r_1} = rV_r + aV_r \cos \phi \tag{16}$$

$$R_2 V_{r_2} = rV_r - aV_r \cos \phi. \tag{17}$$

Doppler Shifts

The approximate Doppler shifted frequency, f_s, reflected from the satellite, will be related to the frequency, f_1, transmitted from station 1 by

$$f_s = f_1\left(1 + \frac{V_{r_1}}{c}\right). \tag{18}$$

The Doppler shifted frequency f_2 received at station 2 will be

$$f_2 = f_s\left(1 + \frac{V_{r_2}}{c}\right). \tag{19}$$

Relating f_2 to f_1,

$$f_2 = f_1\left(1+\frac{V_{r_1}}{c}\right)\left(1+\frac{V_{r_2}}{c}\right) \tag{20}$$

$$f \approx f_1\left(1+\frac{V_{r_1}+V_{r_2}}{c}\right). \tag{21}$$

The net Doppler shift between station 1 and 2 will be

$$\Delta f = (f_2 - f_1) = f_1\left(\frac{V_{r_1}+V_{r_2}}{c}\right). \tag{22}$$

Substituting the previously derived expressions yields

$$\Delta f \approx \frac{f_1 r V_r}{R_1 R_2 c}\left[(R_1+R_2)+(R_2-R_1)\frac{a}{r}\cos\phi\right]. \tag{23}$$

The values of the squares of the ranges may be written

$$R_1^2 = Z_0^2\left[1+\left(\frac{a}{Z_0}\right)^2+\left(\frac{r}{Z_0}\right)^2+\frac{2ar\cos\phi}{Z_0^2}\right] \tag{24}$$

$$R_2^2 = Z_0^2\left[1+\left(\frac{a}{Z_0}\right)^2+\left(\frac{r}{Z_0}\right)^2-\frac{2ar\cos\phi}{Z_0^2}\right]. \tag{25}$$

For the case where $r \overline{<} \frac{1}{5} Z_0$, the ranges may be approximated as

$$R_1^2 = R_2^2 \approx Z_0^2 + a^2. \tag{26}$$

Under this condition

$$\Delta f \approx \frac{\pm 2 f_1 r V_r}{\sqrt{Z_0^2+a^2}\,c} \approx \frac{\pm 2 f_1 r V_r}{\sqrt{2} Z_0 c}$$

$$0 \leq r \leq 300 \text{ nm}, \quad Z_0 \approx a. \tag{27}$$

For the case where

$$\phi = \pm 90°$$

$$R_1^2 = R_2^2 = Z_0^2 + a^2 + r^2 \tag{28}$$

and

$$\Delta f = \frac{2 f_1 r V_r}{\sqrt{3} Z_0 c}$$

$$300 \leq r \leq 1200 \text{ nm max}$$

$$Z_0 \approx a \approx r. \tag{29}$$

For the case where

$$\phi = 0 \text{ or } 180°$$

$$R_1^2 = Z_0^2 + a^2 + r^2 \pm 2ar \tag{30}$$

$$R_2^2 = Z_0^2 + a^2 + r^2 \pm 2ar. \tag{31}$$

Using the assumed values:

$$R_1 \approx \sqrt{5} Z_0 \tag{32}$$

$$R_2 \approx Z_0 \tag{33}$$

and

$$\Delta f \approx \pm \frac{2 f_1 r V_r}{\sqrt{5} Z_0 c}$$

$$300 \leq r \leq 1200 \text{ nm max}$$

$$Z_0 \approx a \approx r. \tag{34}$$

If the origin of time is selected as the origin of r the Doppler shift may be written for the intervals of approximations from the relationship of r, velocity, and time. Grossly approximating the function over the entire interval by (27),

$$\Delta f = \frac{\sqrt{2} f_1 V_r^2 t}{Z_0 c} \tag{35}$$

$$f_1 = 3 \text{ kmc}$$

$$c = 3 \times 10^5 \text{ km/second}$$

$$V_r = 6.66 \text{ km/second}$$

$$Z_0 = 2970 \text{ km}$$

$$t = s$$

$$\Delta f = 280t \text{ cps}$$

$$t_{\max} = \pm 334 \text{ seconds}$$

$$\Delta f_{\max} = 93.5 \text{ kc}. \tag{36}$$

In terms of a relative frequency shift,

$$\frac{\Delta f}{f} = 9.3 \times 10^{-8} t$$

$$\left(\frac{\Delta f}{f}\right)_{\max} = 3.1 \times 10^{-5}. \tag{37}$$

Doppler Bandwidth

The minimum bandwidth which may be used to detect a signal which is changing in time may be approximated as[13]

$$B = \sqrt{\frac{df}{dt}}\,. \tag{38}$$

Based on (36) the theoretical Doppler bandwidth of the receiver would be 53 cps. Assuming the use of a frequency standard which provides precision of one part in 10^9 at 3 kmc, the signal could be defined within this band to ±3 cps.

Relative radial velocity between source and receiver will also influence the bandwidth of the transmitted signal. If B_t is the bandwidth, the apparent bandwidth will be

$$B_a = B_T \frac{\Delta f}{f}\,. \tag{39}$$

Using the maximum relative Doppler frequency shift noted previously for a 4-kc channel, the Doppler compression or expansion will be 0.12 cps. Consequently little error will occur in simply considering the entire bandwidth to be translated with the carrier frequency.

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[13] L. N. Ridenour, "Radar System Engineering," McGraw-Hill Book Co., Inc., New York, N. Y., p. 137; 1947.

The Use of a Passive Spherical Satellite for Communication and Propagation Experiments*

T. H. VEA†, MEMBER, IRE, J. B. DAY†, AND R. T. SMITH†, MEMBER, IRE

Summary—**The use of a passive satellite for over-the-horizon communication links has been proposed by Pierce and Kompfner. To confirm the theoretical work which has been done to date, NASA is planning to launch an aluminum coated, 100-foot-diameter plastic sphere for such a communication relay. In addition to confirming the feasibility of a communication link, the NASA sphere can be instrumental in confirming theoretical work in UHF propagation characteristics. With regard to conducting propagation studies, an essential requirement is that the sphere's scattering properties be described in advance of the experiment. The scattering properties of a metallic sphere (whose radius is very large with respect to the wavelength of the incident radiation) are described in this article. It is shown that the sphere's scattered field pattern is a function of the incident radiation's polarization scheme, and that the choice of any particular polarization scheme depends upon the experiment to be performed.**

In a recent article[1] Pierce and Kompfner discussed the efficacy of over-the-horizon communication by means of forward scatter of UHF radiation by a spherical satellite. An aluminum coated plastic sphere (100 feet in diameter) has been developed by NASA for use as a satellite and will be employed as a reflector in a passive communication link.

The exploration of outer space has created renewed interest in propagation effects at UHF and higher frequencies. The relative freedom from atmospheric and ionospheric effects of UHF and SHF radiation has brought this frequency range into prominence for use in communication with and control of space vehicles. Optimum utilization of tracking systems will require extensive knowledge of propagation phenomena. The spherical satellite will provide a medium by which it will be possible to re-evaluate some of the theoretical expressions which go into the make-up of space link calculations.

Knowledge of the reflection properties of the passive satellite will permit the experimental determination of propagation effects. By using the satellite to scatter energy into regions where the primary radiation is not present, effects such as atmospheric loss, fading margins, refraction, and Faraday rotation can be recorded on a continuous basis. More precise statistical data can then be derived to correlate the changes in effects with elevation angle, sun activity, and, in certain special cases, latitude and azimuth angle. Assuming that the scattering properties of the sphere are defined, the design of the necessary experiments is relatively straightforward.

With an aluminum coating several millimeters thick and under illumination by UHF (or higher frequency) radiation, the NASA sphere's scattering properties are essentially those of a high conductivity, fully metallic sphere. At 300 mc, for instance, an aluminum sphere has a skin depth of 5×10^{-3} cm and a skin effect resistance of 5×10^{-3} ohms. The distributions of scattered energy shown in this article were acquired through solution of the classic equations of Mie[2] for the field components. The exact method by which these solutions were acquired can be found in the Appendix.

When the incident plane (see Fig. 1) is linearly polarized, say along the x axis, its field components are expressible in spherical coordinates as

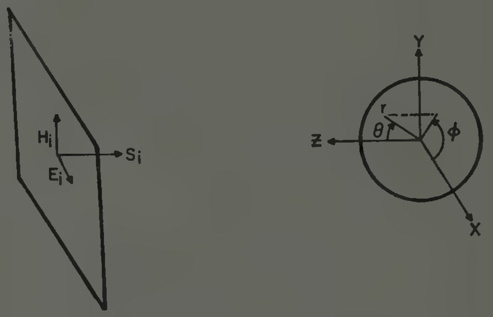

Fig. 1—Incident plane wave.

$$\overline{E}_i = E_0 e^{-i\omega t}(\sin\theta\cos\phi\hat{r} + \cos\theta\cos\phi\hat{\theta} - \sin\phi\hat{\phi})e^{ikr\cos\theta}, \quad (1)$$

$$\overline{H}_i = H_0 e^{-i\omega t}(\sin\theta\sin\phi\hat{r} + \cos\theta\sin\phi\hat{\theta} + \cos\phi\hat{\phi})e^{ikr\cos\theta}; \quad (2)$$

and the radial component of their Poynting vector as

$$S_i = P_0 \cos\theta, \quad (3)$$

where

$$P_0 = \frac{E_0 H_0}{2} = \frac{kE_0^2}{2\omega\mu}.$$

Series solutions, as constructed by Mie and others,[3] are for the scattered field components,

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* Original manuscript received by the IRE, December 2, 1959; revised manuscript received, January 4, 1960.

† Philco Western Development Laboratories, Palo Alto, Calif.

[1] J. R. Pierce and R. Kompfner, "Transoceanic communication by means of satellites," PROC. IRE, vol. 47, pp. 372–380; March, 1959.

[2] J. A. Stratton, "Electromagnetic Theory," McGraw-Hill Book Co., Inc., New York, N. Y., pp. 563–573; 1941.

[3] H. Bateman, "The Mathematical Analysis of Electrical and Optical Wave-Motion," Dover Publications, New York, N. Y.; 1955.

$$\overline{E}_s = E_0 e^{-i\omega t} \sum_{n=1}^{\infty} (i)^n \left[\frac{2n+1}{n(n+1)}\right] \left\{ A_n \left[\frac{\cos\phi}{\sin\theta} P_n'(\cos\theta) h_n(kr)\hat{\theta} - \sin\phi \frac{dP_n'(\cos\theta)}{d\theta} h_n(kr)\hat{\phi}\right] \right.$$
$$- iB_n \left[\frac{n(n+1)}{kr} \cos\phi P_n'(\cos\theta) h_n(kr)\hat{r} + \frac{\cos\phi}{kr} \frac{dP_n'(\cos\theta)}{d\theta} \frac{d}{d(kr)} [krh_n(kr)]\hat{\theta}\right.$$
$$\left.\left. - \frac{\sin\phi}{kr\sin\theta} P_n'(\cos\theta) \frac{d}{d(kr)} [krh_n(kr)]\hat{\phi}\right]\right\}, \tag{4}$$

$$\overline{H}_s = -\frac{kE_0}{\omega\mu} e^{+i\omega t} \sum_{n=1}^{\infty} (i)^n \left[\frac{2n+1}{n(n+1)}\right] B_n \left[-\frac{\sin\phi}{\sin\theta} P_n'(\cos\theta) h_n(kr)\hat{\theta} - \cos\phi \frac{dP_n'(\cos\theta)}{d\theta} h_n(kr)\hat{\phi}\right]$$
$$+ iA_n \left[\frac{n(n+1)}{kr} \sin\phi P_n'(\cos\theta) h_n(kr)\hat{r} - \frac{\sin\phi}{kr} \frac{dP_n'(\cos\theta)}{d\theta} \frac{d}{d(kr)} [krh_n(kr)]\hat{\theta}\right.$$
$$\left.\left. + \frac{\cos\theta}{kr\sin\theta} P_n'(\cos\theta) \frac{d}{d(kr)} [krh_n(kr)]\hat{\phi}\right]\right\}; \tag{5}$$

where

$P_n'(\cos\theta)$ = first degree, nth order associated Lengendre polynomial,
$h_n(kr)$ = nth order Hankel function,
k = propagation constant,
r = distance from center of sphere,
μ = permeability of propagating medium,
ω = radian frequency.

When the sphere has a high conductivity surface and a radius large with respect to wavelength, the boundary conditions for its surface dictates that

$$A_n = -\frac{j_n(ka)}{h_n(ka)}, \quad B_n = -\left[\frac{\dfrac{d}{d(kr)}[krj_n(kr)]}{\dfrac{d}{d(kr)}[krh_n(kr)]}\right]_{r=a}, \tag{6}$$

where

$jn(ka)$ = nth order spherical Bessel function,
a = radius of sphere.

Subsitution of (6) into (4) and (5) yields the following far zone field solutions:

$$\overline{E}_s = \frac{a}{r} E_0 e^{i[k(r-a)-\omega t]} \left\{-\left(\frac{1}{kr}\right)\sin\theta\cos\phi\hat{r}\right.$$
$$\left. - \cos\theta\cos\phi\hat{\theta} + \sin\phi\hat{\phi}\right\} e^{ika\cos\theta}, \tag{7}$$

$$\overline{H}_s = \frac{a}{r}\frac{kE_0}{\omega\mu} e^{i[k(r-a)-\omega t]} \left\{-\left(\frac{1}{kr}\right)\sin\theta\sin\phi\hat{r}\right.$$
$$\left. - \sin\phi\hat{\theta} - \cos\theta\cos\phi\hat{\phi}\right\} e^{ika\cos\theta}. \tag{8}$$

Since the radial component of the far zone field is an inverse function of r^2, it can be neglected in the evaluation of the total far zone Poynting vector,

$$\overline{S}_s = \left(\frac{a}{r}\right)^2 P_0(1 - \sin^2\theta\cos^2\phi)\hat{r}. \tag{9}$$

The pattern of Fig. 2 is a surface of revolution about the x axis and has the same form as the radiation pattern of a quarter wave dipole antenna.

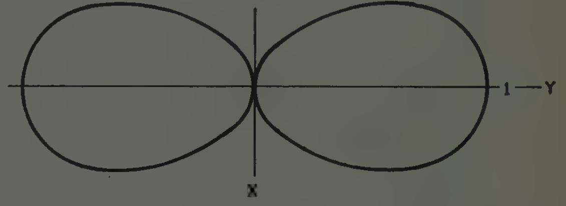

Fig. 2—Scattered energy in sphere's equatorial plane (linear polarization).

If one assumes the incident energy to be linearly polarized along the y axis, the scattered energy distribution is a surface of revolution about the y axis and, in general, the null of the scattered energy distribution can be seen to align itself with the incident polarization vector. When the incident energy is linearly polarized, the scattered field will also be linearly polarized. By control of the spatial orientation of the transmitting and receiving elements, the amount of energy communicated in the link can be maximized (provided that the effects of Faraday rotation can be compensated for).

The field distributions for circular polarization were computed by introducing space quadrature terms and a phase angle parameter β into (1) and (2); hence,

$$\overline{E}_i = E_0 e^{i\omega t}[(\cos\theta\cos\phi + \cos\theta\sin\phi e^{i\beta})\hat{\theta}$$
$$- (\sin\phi - \cos\phi e^{i\beta})\hat{\phi}] e^{ikr\cos\theta}, \tag{10}$$

$$\overline{H}_i = H_0 e^{i\omega t}[(\cos\theta\sin\phi - \cos\theta\cos\phi^{i\beta})\hat{\theta}$$
$$+ (\cos\phi + \sin\phi e^{i\beta})\hat{\phi}] e^{ikr\cos\theta}. \tag{11}$$

The scattered field components were then

$$\overline{E}_s = \frac{a}{r} E_0 e^{i[k(r-a)-\omega t]}\left[-(\cos\theta\cos\phi + \cos\theta\sin\phi e^{i\beta})\hat{\theta} + (\sin\phi - \cos\phi e^{i\beta})\hat{\phi}\right] e^{ika\cos\theta}, \quad (12)$$

$$\overline{H}_s = \frac{a}{r}\frac{kE_0}{\omega\mu} e^{i[k(r-a)-\omega t]}\left[-(\sin\phi - \cos\phi e^{i\beta})\hat{\theta} - (\cos\theta\cos\phi + \cos\theta\sin\phi e^{i\beta})\hat{\phi}\right] e^{ika\cos\theta}, \quad (13)$$

and the far zone Poynting vector,

$$\overline{S}_s = \left(\frac{a}{r}\right)^2 \frac{P_0}{2}(1 + \cos^2\theta - \sin 2\phi \sin^2\theta \cos\beta)\hat{r}. \quad (14)$$

The case of circular polarization results from a substitution of $\beta = \pm\pi/2$ in (12), (13), and (15), whereupon the Poynting vector becomes

$$\overline{S}_s = \left(\frac{a}{r}\right)^2 \frac{P_0}{2}(1 + \cos^2\theta)\hat{r}. \quad (15)$$

The energy distribution of Fig. 3 is a surface of revolution about the z axis. The scattered energy is itself circularly polarized. The values of $\pi/2$ and $-\pi/2$ for β produce an off-set, linearly polarized wave as shown in Fig. 4; these patterns are surfaces of revolution about the $y = -x$ and $y = x$ axes, respectively. For intermediate values of β the patterns will be oriented about the $y = -x$ or $y = x$ axes and will exhibit elliptic polarization.

The free-space loss formula given by Pierce and Kompfner, to be consistent with the distributions of this article, will now read

$$L = \frac{P_t}{P_r} = \frac{\lambda^2 p^4}{A^2 \eta^2 a^2 G}, \quad (16)$$

where

P_t = transmitter power,
P_r = receiver power,
λ = wavelength,
p = geometric mean of the distances between the satellite and the terminals,
A = antenna area (assumed to be the same for both transmitter and receiver),
η = antenna efficiency,
a = radius of sphere,
G = gain of the sphere.

The following form was used for the radar cross section of the sphere:

$$\sigma = 4\pi \frac{S_s}{S_i} = 4\pi a^2 G. \quad (17)$$

The factor G in (17) has a maximum of 0 db for the distributions of this article. An alternate approach would be to compare the scattering of the sphere with that of an isotropic scatterer. The gain of the sphere over an isotropic scatterer is computed in the Appendix and has a maximum of 1.1 db when linear polarization is used. When circular polarization is used the distribution has a maximum of 1.3 db and a minimum of −1.3 db.

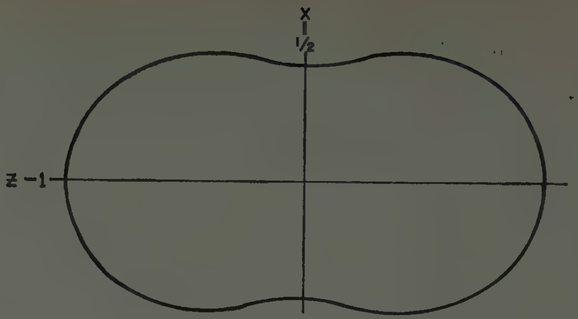

Fig. 3—Scattered energy in sphere's polar plane (circular polarization).

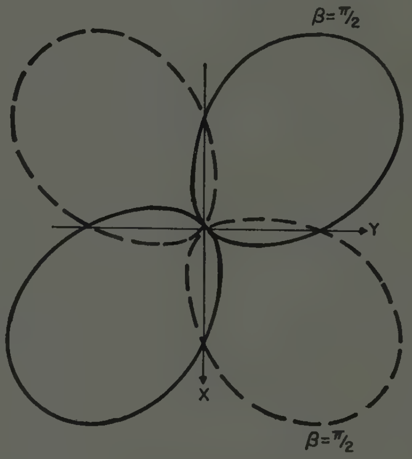

Fig. 4—Switching the field pattern.

Scattering distributions acquired by other investigators[4,5] indicate that the sphere primarily back scatters and forward scatters when the sphere radius is of the order of magnitude of the incident radiation's wavelength. Their published observations were for $a = 3\lambda$ or less, whereas a sphere radius to wavelength parameter of an order of magnitude greater than that just stipulated is pertinent in the UHF region. It would appear that operation of the 100-foot sphere for the experiments mentioned in this article is limited to the region above 100 mc.

The total far zone energy distributions can be obtained from the various field quantities given in this article when both the incident and scattered fields are present. In the case of over-the-horizon communication it is presumed that the incident field energy will not be present, so that expressions (9), (14), and (15) are of principal interest.

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[4] C. I. Beard and V. Twersky, "Propagation through random distributions of spheres," 1958 IRE WESCON CONVENTION RECORD, pt. 1, p. 94.

[5] J. S. Honda, "Scattering of microwaves by figures of revolution," 1957 IRE WESCON CONVENTION RECORD, pt. 1, pp. 151–157.

APPENDIX

A. Solution of the Series

Insertion of (6) into (4) and (5) leads to expressions of the form

$$1)\ h_n(kr)\left[-\frac{j_n(ka)}{h_n(ka)}\right],$$

$$2)\ h_n(kr)\left[-\frac{\dfrac{d}{d(kr)}\,[krj_n(kr)]}{\dfrac{d}{d(kr)}\,[krh_n(kr)]}\right]_{r=a}$$

$$3)\ \frac{d}{d(kr)}\,[krh_n(kr)]\left[-\frac{j_n(ka)}{h_n(ka)}\right],$$

$$4)\ \frac{d}{d(kr)}\,[krh_n(kr)]\left[-\frac{\dfrac{d}{d(kr)}\,[krj_n(kr)]}{\dfrac{d}{d(kr)}\,[krh_n(kr)]}\right]_{r=a};$$

and using

$$h_n(kr)=\frac{e^{ikr}}{i(kr)}\,i^{-n}\sum_{m=0}^{n}\frac{(n+m)!}{m!(n-m)!}\left(\frac{i}{2(kr)}\right)^m, \qquad (18)$$

one obtains

$$1)\ -\left(\frac{a}{r}\right)j_n(ka)e^{ik(r-a)}\pm O\left[\left(\frac{a}{r}\right)^2\right],$$

$$2)\ i\left(\frac{1}{kr}\right)\left[\frac{d}{d(kr)}\,[krj_n(kr)]\right]_{r=a}e^{ik(r-a)}\pm O\left[\frac{a}{kr^2}\right],$$

$$3)\ -ikaj_n(ka)e^{ik(r-a)}\pm O\left[\frac{ka^2}{r}\right],$$

$$4)\ -\left[\frac{d}{d(kr)}\,[krj_n(kr)]\right]_{r=a}e^{ik(r-a)}\pm O\left(\frac{a}{r}\right).$$

Since the far zone fields are of interest, the first terms of the above relations are those used in the solution; hence,

$$\begin{aligned}\overline{E}_s = E_0e^{i[k(r-a)-\omega t]}\sum_{n=1}^{\infty}(i)^n\left[\frac{2n+1}{n(n+1)}\right]\Bigg\{&\left[-\left(\frac{a}{r}\right)\frac{\cos\phi}{\sin\theta}P_n'(\cos\theta)j_n(ka)\hat{\theta}+\left(\frac{a}{r}\right)\sin\phi\,\frac{dP_n'(\cos\theta)}{d\theta}j_n(ka)\hat{\phi}\right]\\ &-i\left[i\,\frac{n(n+1)}{(kr)^2}\cos\phi P_n'(\cos\theta)j_n(ka)\hat{r}-\frac{\cos\phi}{kr}\,\frac{dP_n'(\cos\theta)}{d\theta}\left[\frac{d}{d(kr)}\,[krj_n(kr)]\right]_{r=a}\hat{\theta}\right.\\ &\left.+\frac{\sin\phi}{kr\sin\theta}P_n'(\cos\theta)\left[\frac{d}{d(kr)}\,[krj_n(kr)]\right]_{r=a}\hat{\phi}\right]\Bigg\}, \end{aligned} \qquad (19)$$

$$\begin{aligned}\overline{H}_s = -\frac{kE_0}{\omega\mu}\,e^{i[k(r-a)-\omega t]}\sum_{n=1}^{\infty}(i)^n\left[\frac{2n+1}{n(n+1)}\right]\Bigg\{&\left[-i\,\frac{\sin\phi}{(kr)\,\sin\theta}\,P_n'(\cos\theta)\left[\frac{d}{d(kr)}\,[krj_n(kr)]\right]_{r=a}\hat{\theta}\right.\\ &\left.-i\,\frac{\cos\phi}{kr}\,\frac{dP_n'(\cos\theta)}{d\theta}\left[\frac{d}{d(kr)}\,[krj_n(kr)]\right]_{r=a}\hat{\phi}\right]+i\left[\frac{n(n+1)}{kr}\left(\frac{a}{r}\right)\sin\phi P_n'(\cos\theta)j_n(ka)\hat{r}\right.\\ &\left.-i\left(\frac{a}{r}\right)\sin\phi\,\frac{dP_n'(\cos\theta)}{d\theta}j_n(ka)\hat{\theta}-i\left(\frac{a}{r}\right)\frac{\cos\phi}{\sin\theta}\,P_n'(\cos\theta)j_n(ka)\hat{\phi}\right]\Bigg]. \end{aligned} \qquad (20)$$

Collecting terms of the various coordinates

$$\begin{aligned}\overline{E}_s = E_0\left(\frac{a}{r}\right)e^{i[k(r-a)-\omega t]}\Bigg\{&\sum_{n=1}^{\infty}(i)^n(2n+1)\left(\frac{1}{kr}\right)\frac{\cos\phi}{ka}\,P_n'(\cos\theta)j_n(ka)\hat{r}\\ &-\sum_{n=1}^{\infty}(i)^n\left[\frac{2n+1}{n(n+1)}\right]\left[\frac{\cos\phi}{\sin\theta}\,P_n'(\cos\theta)\,j_n(ka)-i\,\frac{\cos\phi}{ka}\,\frac{dP_n'(\cos\theta)}{d\theta}\left[\frac{d}{d(kr)}\,[krj_n(kr)]\right]_{r=a}\right]\hat{\theta}\\ &+\sum_{n=1}^{\infty}(i)^n\left[\frac{2n+1}{n(n+1)}\right]\left[\sin\phi\,\frac{dP_n'(\cos\theta)}{d\theta}j_n(ka)-i\,\frac{\sin\theta}{ka\sin\theta}\,P_n'(\cos\theta)\left(\frac{d}{d(kr)}\,[krj_n(kr)]\right)_{r=a}\right]\hat{\phi}\Bigg\}, \end{aligned} \qquad (21)$$

$$\begin{aligned}H_s = \frac{kE_0}{\omega\mu}\left(\frac{a}{r}\right)e^{i[k(r-a)-\omega t]}\Bigg\{&\sum_{n=1}^{\infty}(i)^n(2n+1)\left(\frac{1}{kr}\right)\frac{\sin\phi}{ka}\,P_n'(\cos\theta)j_n(ka)\hat{r}\\ &-\sum_{n=1}^{\infty}(i)^n\left[\frac{2n+1}{n(n+1)}\right]\left[\sin\phi\,\frac{dP_n'(\cos\theta)}{d\theta}j_n(ka)+i\frac{\sin\phi}{ka\sin\theta}\,P_n'(\cos\theta)\left[\frac{d}{d(kr)}\,[krj_n(kr)]\right]_{r=a}\right]\hat{\theta}\\ &+\sum_{n=1}^{\infty}(i)^n\left[\frac{2n+1}{n(n+1)}\right]\left[\frac{\cos\phi}{\sin\theta}P_n'(\cos\theta)j_n(ka)-i\,\frac{\cos\phi}{ka}\,\frac{dP_n'(\cos\theta)}{d\theta}\left[\frac{d}{d(kr)}\,[krj_n(kr)]\right]_{r=a}\right]\hat{\phi}\Bigg\}. \end{aligned} \qquad (22)$$

The series that comprise (21) and (22) can be identified as those of the type used to set up the incident fields[2] or identified by a technique described in a recent article;[6] hence (7) and (8) have the form

$$\overline{E}_s = \frac{a}{r} E_0 e^{i[k(r-a)-\omega t]} \left\{ -\left(\frac{1}{kr}\right) \sin\theta \cos\phi \hat{r} - \cos\theta\cos\phi\hat{\theta} + \sin\phi\hat{\phi} \right\} e^{ika\cos\theta},$$

$$\overline{H}_s = \frac{a}{r} \frac{kE_0}{\omega\mu} e^{i[k(r-a)-\omega t]} \left\{ -\left(\frac{1}{kr}\right) \sin\theta \sin\phi \hat{r} - \sin\phi\hat{\theta} - \cos\theta\cos\phi\hat{\phi} \right\} e^{ika\cos\theta}.$$

B. The Poynting Vector

Since the radial component of the far zone fields was neglected, the total Poynting vector has the following form:

$$\overline{S}_s = \tfrac{1}{2} \operatorname{Re} E_\theta H_\phi^* - E_\phi H_\theta^*]\hat{r}.$$

For the fields calculated in *A* of the Appendix,

$$\overline{S}_s = \left(\frac{a}{r}\right)^2 \frac{kE_0}{2\omega\mu} [(\sin\theta\cos\phi)^2 + (\sin\phi)^2]\hat{r},$$

$$\overline{S}_s = \left(\frac{a}{r}\right)^2 P_0[1 - \sin^2\theta\cos^2\phi]\hat{r}.$$

C. Equivalent Gain of the Sphere

The equivalent gain of the sphere over an isotropic scatterer was computed from the pattern volume. For linear polarization,

$$V_L = 8\int_0^{\pi/2}\int_0^{\pi/2}\int_0^{\cos^2\phi} r^2 \cos\phi \, dr\,d\phi\,d\theta,$$

$$V_L = \frac{64\pi}{105}$$

from which the radius of the equivalent sphere is computed to be 0.77; hence,

$$G_L = \frac{1}{0.77} = 1.30.$$

For the case of circular polarization,

$$V_c = 8\int_0^{\pi/2}\int_0^{\pi/2}\int_0^{1/2(1+\cos^2\theta)} r^2 \sin\theta \, dr\,d\theta\,d\phi,$$

$$V_c = \frac{16\pi}{35}$$

and the radius of the equivalent sphere was computed to be 0.70; hence

$$G_c = \frac{1}{0.7} = 1.43.$$

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[6] T. Vea, "Generating functions and the summation of infinite series," to be published in Proc. IRE.

Project SCORE*

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Summary—This paper is designed to provide a review and summary of the over-all program of development, construction and operation by the U. S. Army Signal Research and Development Laboratory of the first satellite communication system placed in orbit on December 18, 1958. This effort was part of Project SCORE (Signal Communication by Orbiting Relay Experiment) directed by ARPA. Basically, Project SCORE consisted of designing, constructing, and conducting an actual flight experiment of a delayed repeater type satellite package using an actual stripped down Atlas missile as the satellite container. The major points to be covered are summarized in the following paragraphs.

A brief historical résumé of the project will cover the preliminary discussions regarding the feasibility of orbiting an Atlas ICBM; the vehicle and weight limitation affecting preliminary system design; mechanical and electrical test problems and ground station installation and on-site training.

The system description will cover orbit data; ground station site location and control; airborne equipment components such as antennas, temperature control and communication electronics; general considerations of system design and design problems; ground station equipment including recording and test equipment and associated ground communication equipment.

The system operation portion of the paper will cover a brief summary of the launch into orbit and actual life of the communications system; actual operating procedures used to achieve maximum benefits from the experiment; a summary of test data obtained with chart and graph presentations; a discussion of test results and the factors affecting the evaluation of test results; and selective comments on equipment reliability.

The final part of the paper will contain conclusions to be made from the equipment and the impact of these conclusions on the planning for future communications satellite systems.

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STARTING with the development of radio communication equipment suitable for operational use in the VHF and UHF frequencies, system designers and operating personnel alike have been concerned with the problems imposed by line-of-site propagation at these frequencies. Equipment operating on even higher frequencies with corresponding greater communication channel capacity is being developed and placed in operational status. However, no reliable and economic means of providing single-link, wide-band communication facilities over intracontinental and intercontinental distances yet exists.

To fill this communication gap, extensive networks of coaxial line and microwave radio relay systems have been developed and installed in the continental United States during the past decade. The logistics problems attendant upon the installation and maintenance of these systems in the confines of the United States are tremendous, even with a relatively favorable environment. Applying the same techniques to a fluid military situation aggravates the difficulties to produce an almost impossible situation. The economics of the problem of providing multichannel communication under these conditions are such that only the pressure of military necessity justifies the cost. Even in commercial practice, the installation of radio relay and coaxial cable facilities is limited to the land masses of the eastern and western hemispheres. There are no means available to cross the reaches of the Atlantic and Pacific Oceans with economically sound and reliable radio relay or coaxial facilities capable of transmitting, for example, real time television program material.

Many schemes have been proposed, such as circling aircraft, blimps or dirigibles, land stations installed on the islands which dot the subarctic borders of the Atlantic and Pacific Oceans, anchored platforms and cables with sealed-in repeaters. Each of these, except the last, in one form or another has developed flaws, either logistic or technical, which makes them extremely unattractive, if not impossible. The last has been accomplished in recent years for multichannel telephone and slowed-down television; however, it is quite vulnerable to earthquakes, tides and Soviet trawlers.

With the successful launching of man-made satellites, their possibilities as elements in world-wide transmission systems were the subject of considerable speculation. A satellite, quite obviously, has many attractions as a vehicle for an unattended radio relay repeater. It requires no fuel to maintain it on its appointed course. Once in orbit, its position can be predicted with great accuracy and by selection of orbits, with no more than two relay points, any two locations on earth can be linked.

Attractive as these possibilities are for commercial communication systems, they become almost irresistible for world-wide military networks. At the present time the satellite station is practically impossible to intercept. It requires no friendly real estate to house equipment and, powered either by atomic energy or solar radiation, needs no logistic support. Operating as a delayed repeater, a satellite link could give a military system a considerable antijamming advantage.

Naturally, there are some disadvantages. The electronic equipment must be rugged enough to withstand the hazards of launching, one hundred per cent reliable for a time at least sufficient to amortize its initial cost and launching expenses, and capable of withstanding the extreme conditions existing in outer space. None of these appear to be insurmountable at the present time.

It was in an environment of such considerations that the Advanced Research Projects Agency, jointly with the Army and the Air Force, undertook Project SCORE. The object of this project was to place in orbit an eighty-foot long Atlas missile and to use this as a platform for a communication system capable of spanning intercontinental distances. The ultimate goal was to demonstrate dramatically the feasibility of such a system and to explore some of the technical and operational problems that would attend a military satellite communication system. The communications portion of the project was approved and assigned to the U. S. Army Signal Research and Development Laboratory late in July, 1958. The launching date was set for early November, 1958, which required that system and equipment design and installation in the vehicles had to be completed by the middle of October to avoid interference with prelaunch checkouts. Ground stations were required to be installed and operating crews trained by November 1, 1958. This was admittedly an extremely short deadline within which to accomplish project SCORE.

The plans were to install two complete communication packages in the Atlas missile. These each would include a receiver, transmitter, control unit, battery supply tape recorder, and tracking beacon. Ground equipments installed in standard army type V-51 vans with associated support vehicles were to be located at Fort MacArthur, Calif., Fort Huachuca, Ariz., Fort Sam Houston, Tex., and Fort Stewart, Ga. This equipment would include a Quad helix tracking antenna mounted on searchlight base, receiving, transmitting and control equipment along with appropriate recording, telephone and teletype terminal equipment. In addition, the California station would have a direction finder to assist in initial pickup and tracking of the satellite. All ground stations were to be linked by both telephone and HF radio to a system control center at the Deal Area of the Signal Corps Laboratory in New Jersey.

Preliminary calculations indicated that an elliptical orbit with an apogee of between 500 and 800 miles and a perigee of approximately 100 miles might be achieved. The launching was to be from the Atlantic Missile Range at Cape Canaveral with an inclined orbit of 30°. Fig. 1 illustrates the orbit geometry and the location of the four operating ground stations.

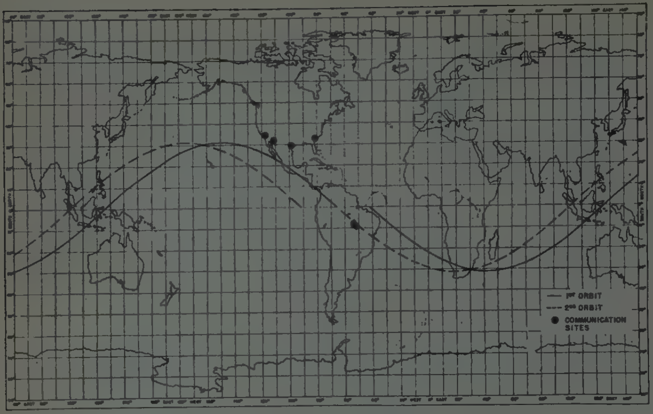

Fig. 1—SCORE satellite orbit.

The design of the system was based on providing two modes of operation; as a delayed repeater and as a real time active satellite repeater. In the delayed repeater mode, the satellite, upon reception of a suitable command signal from a ground station would record information transmitted to it. Upon reception of a different command signal, it would then transmit back to the ground station originating the command the information previously stored. Within the limitation of the storage capacity in the satellite, the information transmission rate, and hence the bandwidth, of the system and the time in view over each of the ground stations, a delayed repeater can provide intercontinental "store and forward" message transmission service.

The second mode of operation, that of a real time repeater, was obtained by the use of yet another command signal which activated the satellite as a radio relay repeater station with the recording mechanism bypassed. Here, the transmission capability of the system is limited by the time the satellite is in view of both stations simultaneously and the system bandwidth. The locations of the four ground stations were based, in part, on providing station separations which would permit testing of the real time capability over varying distances.

The system parameters which were incorporated in the satellite and ground equipments are indicated in Table I. VHF frequencies were used to minimize the effects of cosmic noise and ionospheric propagation while still permitting the use of sensitive, transistorized receiving equipment in the satellite. The power levels employed were consistent with what could be obtained in the time frame of the program. The bandwidths selected were the minimum consistent with frequency stability, Doppler frequency variations, maximum audio frequency and carrier frequency deviation. Allowance was made for ±4-kc variations caused by Doppler shift and ±5 kc for the first-order sidebands of the modulation with the deviation ratio limited to 1.0 at 5 kc. The frequency stability of the satellite components was within ±0.005 per cent. Consistent with the expected apogee of 500 to 800 miles, the system design was based on 1000-miles slant range. Power balance calculations using this value resulted in the fade margins indicated, including allowances for miscellaneous losses in transmission lines, duplexers, filters and ring isolators.

TABLE I

SCORE System Parameters

	Satellite	Ground
Transmit Frequency	132 mc	150 mc
Receive Frequency	150 mc	132 mc
RF Power Output	8 watts	1000 watts
Noise Figure	10 db	6 db
IF Bandwidth	40 kc	40 kc
Audio Bandwidth	0.3–5.0 kc	0.3–5.0 kc
Antenna Gain	0 db	16 db @150 mc
FM Threshold	10 db	10 db
Fade Margin	39 db	19 db

The satellite elements of the system were designed with due consideration for simplicity, the capability to withstand the mechanical environment during powered flight, an adequate thermal design to maintain the electronic components within reasonable temperature limits while in orbit, and redundancy to improve the likelihood of a successful communications experiment. Two complete systems, using slightly different radio frequencies, were installed in the Atlas missile to provide the redundant feature. Table II summarizes the size, weight and power drain for the items comprising the over-all satellite system.

TABLE II

SCORE Component Characteristics

Components	Weight (Pounds)	Size (Cubic Inches)	Power Drain (Watts)
Receiver	0.75	39	0.24
Transmitter	2.5	101	51.3
Beacon	1.0	27	0.36
Control Unit	0.75	37	3.0*
Recorder	3.25	99	1.4
DC-DC Converter	1.0	19.2	—
Battery	21.0	175	—
Mount and Hardware	14.5	—	—
Antenna System	54.0	—	—

* Relay mode only.

The major design areas for the satellite system can be broken down into the following areas which are described below.

The major elements of each of the satellite systems are shown in Fig. 2 as well as the over-all configuration. Fig. 3 indicates the interconnection between the major elements. As shown in Fig. 3, separate antennas were used for the communication receiver and transmitter with a beacon transmitter diplexed with the higher powered communication transmitter.

The communications receiver was a double superheterodyne frequency-modulated all-transistor receiver, a modified version of a commercial type. Considerable improvement in sensitivity was attained by the use of selected transistors in an RF amplifier stage which was included; noise figures in the 10-db to 13-db range were

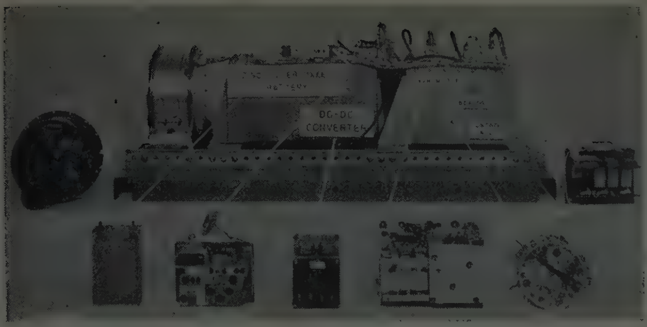

Fig. 2—SCORE system satellite components.

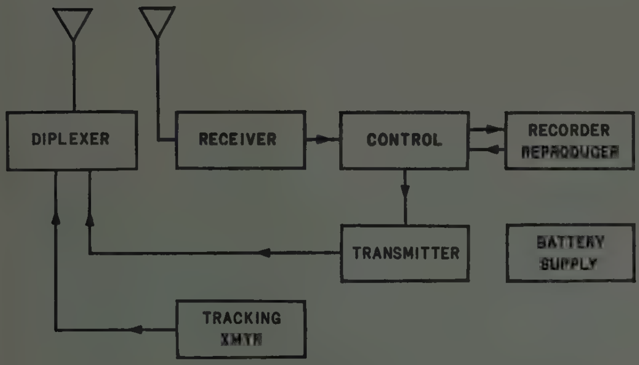

Fig. 3—SCORE satellite interconnection diagram.

realized. Although the receiver power drain was only 0.24 watt, it represented a considerable expenditure of energy over long periods of time if it were permitted to operate continuously. To conserve the limited battery supply, a battery saver circuit was included in the receiver design which, during standby periods between interrogations, turned the receiver on for a period of $\frac{1}{4}$ second each $2\frac{1}{2}$-second period, thereby reducing the power requirement by a factor of 10. This substantial saving in power drain was thereby achieved without introducing appreciable delay in detecting command signals.

The transmitter used vacuum tubes throughout and provided 8-watts frequency-modulated power output. The circuitry used was conventional, in keeping with the urgency of the development cycle. Plate and screen voltages for the transmitter were obtained from a transistorized dc to dc converter operating from the battery power supply. A relatively high conversion efficiency of 82 per cent was obtained with the converter operating with a ripple frequency of 2500 cps.

The control unit consisted of transistorized switching circuits activated by special telemetry type notch filters at frequencies of 3000, 3900 and 5400 cps. These tones corresponded to the record, playback and radio relay modes, respectively, and caused the appropriate connections to be made between the transmitter, receiver and tape recorder. The 5400-cps tone had an alternate capability of disabling the record and playback modes and caused the system to revert to a standby status. This was included to provide a measure of control of the satellite system by the ground stations in the event that the automatic cycling sequence incorporated in the satellite design might fail.

The recorder-reproducer was a miniature type originally designed and built at the Signal Corps Laboratory for a meteorological satellite, but modified for application to the SCORE project. The electronic circuitry including erase, playback and record functions was completely transistorized; the tape drive mechanism was pressurized for reliable operation of the dc motor in the space environment. An endless loop of 75 feet of 1-mil mylar tape, operating with a tape speed of $3\frac{3}{4}$ inches per second (ips) provided 4 minutes of recording or playback time with a 300- to 5000-cps bandwidth. A $\frac{1}{2}$-inch metallic coated section of the tape which momentarily completed an electrical circuit indicated the stopping point on the endless loop. This contact caused the control unit to revert automatically to a standby status at the completion of a record or playback cycle.

The nominal 108-mc tracking beacon used was a modified version of the transistorized type used in Explorer I. One channel of FM-AM telemetry was included in the design to indicate temperature variations in either the missile pod cover or one of the electronic packages. Output power was a continuous 30 milliwatts.

Based on the expected life of the satellite of 21 days, a chemical battery system rather than a costlier and heavier solar cell and rechargeable nickel cadmium battery system was used. Zinc-silver oxide batteries were chosen because of their high capacity (watt hours per pound) and ability to handle occasional peak loads, such as the communication transmitter.

The antennas employed consisted of two simple slot antennas for reception and two for transmission, a receiving and a transmitting antenna being located on each instrumentation pod of the missile. A typical slot antenna is shown in Fig. 4 which was fabricated from fiberglass honeycomb sandwich material with the conducting surfaces applied by means of an aluminum metal spray process.

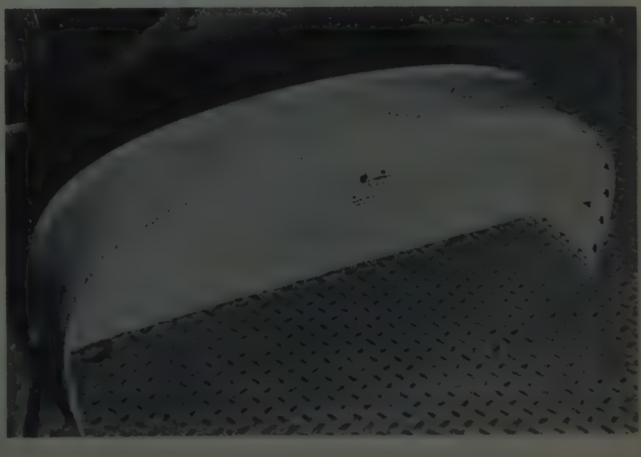

Fig. 4—Slot antenna used with SCORE satellite.

Fig. 5 indicates the manner in which the various components were interconnected; band-pass filters and ring couplers being used to isolate the effects of like units from each other while still providing a low-pass path connection to the antennas located on opposite pods of the missile. The separate transmitting and receiving antennas provided the additional decoupling necessary to eliminate desensitization of the receiving equipment.

The antenna radiation pattern attained was essentially that of a multiwavelength doublet with the attendant deep nulls. The effect of the missile tumbling in its orbital flight was to introduce fades of up to 15 db in the transmission path. The fade margins indicated in Table I were made as large as possible to overcome this effect.

Fig. 6 indicates the over-all configuration of each system mounted on rails on the Atlas missile. The electronic components were coated, where circuit configurations permitted, with a conformal resin coating. All of the major assemblies were further encapsulated in an isocyanate foam, and poured into place within lightweight, polished aluminum canisters. Printed wiring was further used in most units. The finally assembled units, in their canisters, were attached to the mounting rails through isocyanate foam and laminated fiberglass mounts, with the exception of the recorder where solid Kel-F plastic mounts were used. These mounts provided a measure of vibration isolation but, more important, thermally isolated the electronic equipment from the skin of the missile. The over-all configuration withstood accelerations of 10 *g* from 20 to 2000 cps and $\frac{1}{4}$-inch excursions from 5 to 20 cps.

The thermal design presented unique problems because of the difficulty in providing an adequate thermal environment in the vicinity of the electronic components. A range of +40°F to +120°F was the largest useable, if system performance were to be met. The approach finally decided upon was to use highly polished surfaces on the communication components while the surface of the metallic pod covers were iridited. This was expected to produce a mean temperature of approximately 90°F within the individual packages. The expected low duty cycle of the communication transmitter was not expected to cause an appreciable temperature rise within its canister. Since the temperature compensation problem was not straightforward and was severely limited by payload and development time considerations, the results obtained were not unexpected.

The ground stations were designed to be operated from transportable vans and trucks and included certain minimum requirements. In addition to the radio terminal equipment, including transmitters, receivers, telegraph and telephone terminal equipment, as shown in Fig. 7, it was necessary to provide means for tracking the orbiting satellite with a high gain antenna, recording pertinent signal level and communication data, and automatically sequencing the ground equipment for either transmission or reception with the satellite.

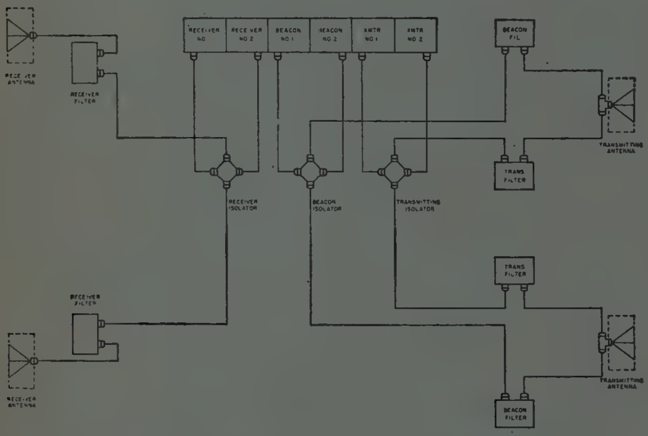
RECEIVER NO
RECEIVER NO 2
BEACON NO 1
BEACON NO 2
XMTR NO 1
XMTR NO 2
BEACON FIL
RECEIVER ANTENNA
RECEIVER FILTER
RECEIVER ISOLATOR
BEACON ISOLATOR
TRANSMITTING ISOLATOR
TRANSMITTING ANTENNA
TRANS FILTER
TRANS FILTER
RECEIVER FILTER
RECEIVER ANTENNA
TRANSMITTING ANTENNA
BEACON FILTER

Fig. 5—SCORE antenna connection diagram.

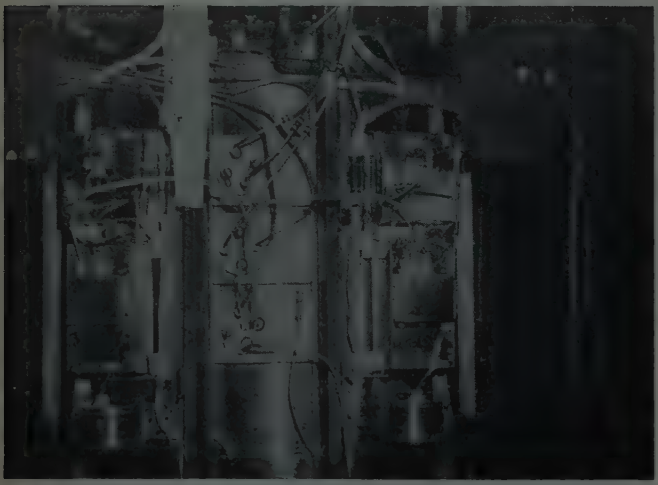

Fig. 6—SCORE satellite components mounted on Atlas 10B missile.

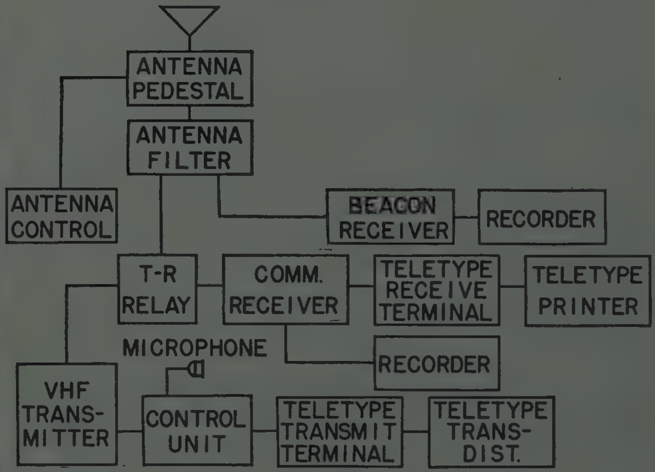
ANTENNA PEDESTAL
ANTENNA FILTER
ANTENNA CONTROL
BEACON RECEIVER
RECORDER
T-R RELAY
COMM. RECEIVER
TELETYPE RECEIVE TERMINAL
TELETYPE PRINTER
MICROPHONE
RECORDER
VHF TRANS-MITTER
CONTROL UNIT
TELETYPE TRANSMIT TERMINAL
TELETYPE TRANS-DIST.

Fig. 7—SCORE ground station interconnection diagram.

The VHF radio terminals were modified commercial types with characteristics as listed in Table I. Backups were provided for all major items in the communications chain with both a 250-watt and a 1000-watt transmitter, for example. These units occupy the center of Fig. 8, an interior view of the operations van. The teletype multiplex equipment was capable of either seven-channel, frequency division multiplex operation or single-channel, 850-cps frequency shift operation. Standard military teletype sets were used for receiving, transmitting and preparing messages on perforated tape.

The control unit shown in Fig. 9 not only originated the command tones transmitted to the satellite, but also automatically accomplished the switching of the receivers, transmitters, teletype machines and other equipment in the proper sequence. At the operators' position, a patch panel was also provided which permitted changing the operating sequence or type of service at a single location.

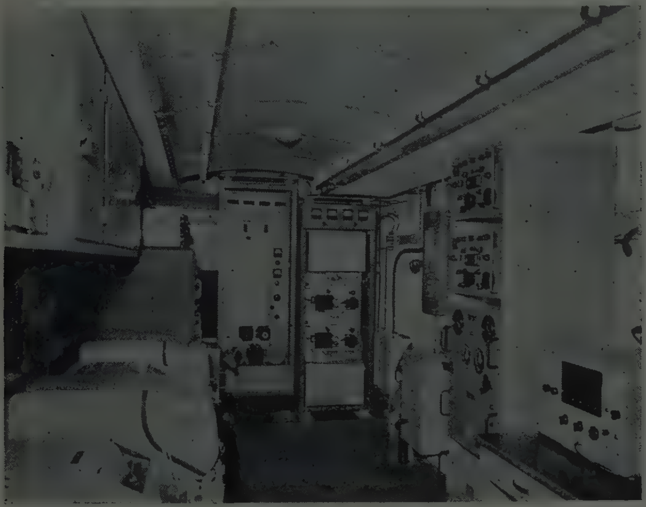

Fig. 8—Interior view of SCORE ground station operations van.

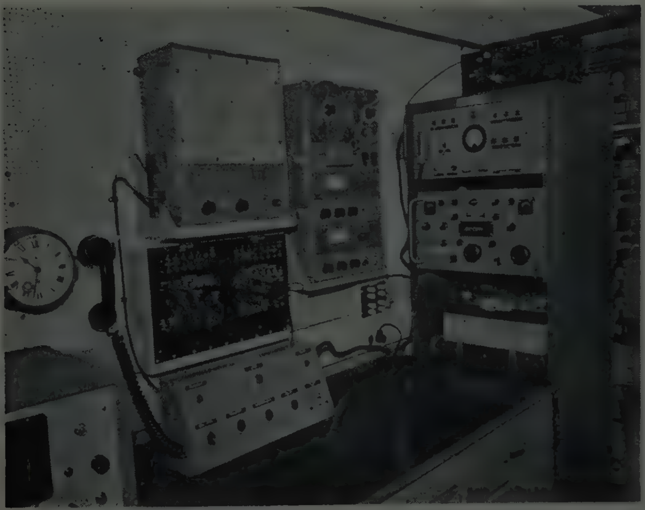

Fig. 9—Operator's console in SCORE ground station.

Recording devices, including magnetic tape recorders and pen recorders, were also supplied to provide permanent records of such things as signal level variations and communication traffic.

The antenna system used (see Fig. 10) was a circularly polarized, quad-helices array with a screen reflector mounted on a modified searchlight pedestal. The gain provided varied from 10 db at the 108-mc tracking frequency to 16 db at the 150-mc communication frequency. Coverage of 360° in azimuth and 90° in elevation was controlled at all but the California station by an operator maximizing the 108-mc signal reception with the antenna positioning controls located near the tracking receivers. At the California station, the azimuth control was slaved to the alidade of an experimental direction-finding equipment while the elevation control was manually varied. Even this partial automation resulted in a marked improvement in tracking accuracy and system performance.

The successful launching of the Atlas missile 10B on December 18, 1958 and the subsequent operation of the communication system installed therein almost instantly assumed importance as an achievement of national and international significance. This fact interfered with the conduct of the technical portions of the project to such an extent that the orderly collection and analysis of data were somewhat neglected for the first week of operation. Therefore, some important clues that might have explained some discrepancies later observed were lost. Qualitatively, the practical operation of a satellite radio relay system capable of spanning intercontinental distances either in real or delayed time was demonstrated conclusively until battery failure occurred on December 30, 1958. For example, communication traffic, both voice, single-channel teletype and multichannel teletype, was carried on a delayed repeater basis in many successive orbits around the earth and was

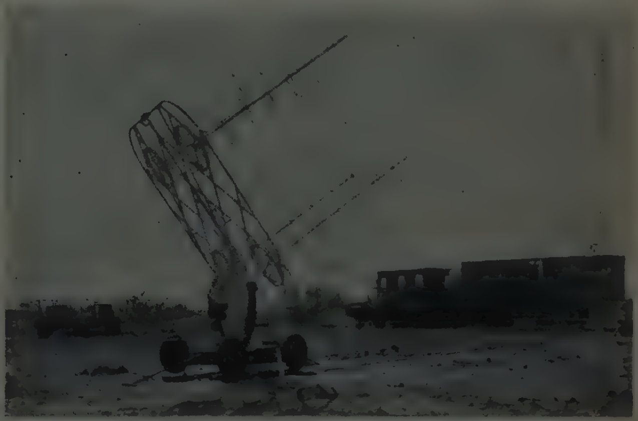

Fig. 10—Exterior view of SCORE ground station showing tracking and communication antenna.

made available individually to the four ground stations at their command.

The ground stations interrogated, received and recorded, for simulated distribution, a total of 78 messages for a total of 5 hours, 12 minutes communication time. Of these, 28 separate messages had been placed in the satellite recorder by ground stations. The remainder were multiple interrogations of previously recorded material. A total of 1 hour and 52 minutes operation of the satellite in the recording mode was involved for these 28 loadings. Eleven real time relays from California to Georgia, a distance of over 3000 miles, were successfully made for a total of 43 minutes of satellite operation. In several instances, unauthorized fortuitous recordings from unidentified stations, with apparently random distribution in the Eastern hemisphere, were received during interrogations by both the California and Georgia stations. In all probability, many fortuitous interrogations of the satellite also were made but there were no means of determining the number of these. This does indicate that a simple tone keyed command system is highly inadequate for future satellite applications.

One set of equipment failed because the tape recorder became inoperative on the first orbit allowing the transmitter to operate continuously and discharge the battery. The other equipment operated exactly as planned other than with somewhat smaller fade margins than had been expected. This satisfactory operation was obtained in spite of the fact that internal temperatures appeared to rise to over 20°F higher than the maximum temperature of 120°F for which the equipment was designed.

This system, which was assembled in record time, demonstrated a world-wide single-voice channel communication capability, over intercontinental distances, in real time, and to and from any point on earth by delayed (store and forward) repeater techniques.

There is no insurmountable barrier to the expansion of the principles demonstrated by this project to the dimensions necessary for an economically practical, world-wide communication system. The communication equipments developments required are well within the state of the art, namely, wide-band transmitters and receivers, a recording facility of complementary capacity and a secure command control system, all capable of providing unattended operation for a year or more by using redundancy and other techniques. It is beyond the purpose of this paper to analyze the bandwidth and time dimensions necessary for economic feasibility, but, in all likelihood, the electronic equipment could be available in about the same time frame as the rocket power necessary to place such a satellite into a proper orbit.

A Broad-Band Spherical Satellite Antenna*

HENRY B. RIBLET†, SENIOR MEMBER, IRE

Summary—**The design criteria and results of the design of a broad-band spherical antenna are discussed. The parameters of design and their effect on radiation patterns are briefly investigated. The particular antenna was developed from an equiangular spiral slot plotted on a plane and then projected on the surface of a sphere. Attention is given to problems of isolation and matching when the antenna is used for multifrequency operation. The particular antenna discussed is being used in the TRANSIT satellite program.**

INTRODUCTION

THE need for a broad-band antenna, covering a frequency of at least four to one, is dictated by the use of the four harmonically related frequencies 54, 108, 162, and 216 megacycles in the TRANSIT satellite. These four frequencies are used to obtain data pertaining to the ionosphere refraction observed during Doppler measurements. In addition to the wide range of frequencies, the desirability of radiation patterns as near isotropic as possible led to the design of the equiangular spiral slot antennas projected onto the surface of a sphere. This antenna provides radiation patterns which have nulls no deeper than 10–15 db, at any angle around the sphere, except the one produced by cross polarization at the equator.

ANTENNA CONFIGURATION

Unless the satellite is stabilized with respect to the earth, the variations in attitude caused by spin, tumbling, and other changes in space orientation will require the radiation pattern to have a uniform field intensity in all directions, *i.e.*, isotropic, in order to minimize the signal strength variations at the receiver. Because it is next to impossible to obtain a practical isotropic antenna, the specifications for the TRANSIT satellite antenna call for peak-to-peak field intensity variations of not more than 10 db within the spherical coordinate system for all four frequencies. Other requirements make the use of protruding dipole elements mechanically difficult. With few exceptions, the flush-mounted spiral slot satisfied these specifications.

The spherical shape of the TRANSIT satellite was chosen initially to obtain thermal balance regardless of attitude; however, the same property of the sphere, having only one linear dimension, the circumference, aids greatly in obtaining a nearly uniform radiation pattern. To fulfill the broad-band frequency requirement, the properties of an equiangular (logarithmic) spiral slot was utilized. The shape and dimensions of a spiral slot antenna are specified entirely by angles. When this is true, its electrical dimension is constant with wavelength; hence, its performance is "broad-band." These broad-band characteristics give uniform beam widths and impedances over frequency ranges up to 10 to 1. Other familiar examples of broad-band antennas are the biconical horn and the disk-cone antenna.

As shown in Fig. 1, the equiangular (*i.e.*, logarithmic) spiral has the property that the angle A between the tangent to the curve and the radius vector at any point on the curve is a constant. Angle A must be larger than 90° to qualify as a spiral. The spiral can be defined by

$$\rho = Ke^{a\theta}.$$

This equation describes the curve that forms one edge of the spiral slot or the conductor in the case of its complementary form, a conducting spiral element. The other edge of the spiral can then be defined by

$$\rho' = Ke^{a(\theta-\alpha)},$$

where the angle α is lag angle between equal radius vectors as shown in Fig. 2.

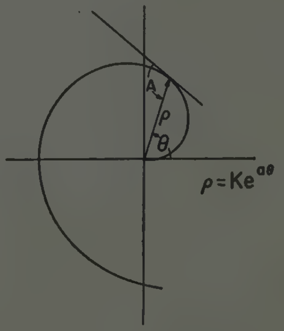

Fig. 1—Construction of the equiangular spiral.

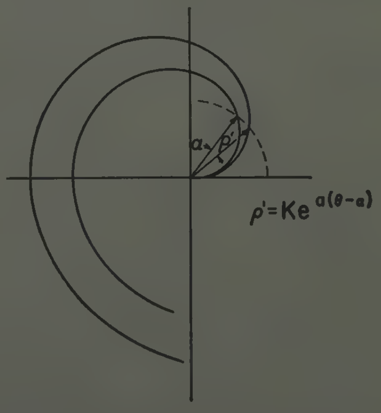

Fig. 2—Construction of two spirals with a difference angle α.

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The slot width is the distance between ρ and ρ' at equal angles of θ. A double spiral slot can be constructed by plotting two spirals 180° with respect to each other as shown in Fig. 3.

The double spiral slot was projected on the surface of each hemisphere to complete the antenna as shown in Fig. 4.

RADIATION PATTERNS

The nearly uniform radiation patterns are obtained by controlling the current and phase distributions over the entire surface of the sphere. With separate balanced spiral slots on each hemisphere one has a few parameters which can be adjusted to achieve optimum performance. These parameters are the screw sense of the spiral, the feed point phase, and the rotational position of one hemisphere with respect to the other. There are several combinations of these parameters which will produce varied radiation patterns.

In the case of the satellite, the antenna must be designed to give as near a uniform radiation field as possible. It is necessary, therefore, to adjust these parameters so that in any plane through the poles of the sphere the current distribution is in phase for any set of symmetrical points. If the hemispheres are isolated by a metal conducting disk through the equator, the radiation from each hemisphere in the direction of its pole is essentially independent since the other hemisphere is completely shadowed. Now as the point of radiation moves toward the equator, radiation components from the other hemisphere become increasingly contributing, until the field from the equator has equal components of radiation from each hemisphere. If these components can be made to be in phase for all longitudinal angles around the equator, they will add to fill in the null of revolution to the maximum extent. Fig. 5(a) shows that for a fixed feed point phase and for spirals wound in the opposite sense, the electric field across the slot will be in phase at all points around the equator providing the two spirals are lined up in a rotational sense so that the ends and the beginning of the spirals are on the same longitudinal angle. By the same technique, Fig. 5(b) shows that for spirals wound in the same sense, the radiation will be in phase at $\phi=0$ and out of phase at $\phi=90°$.

Fig. 6 shows the evolution of the radiation patterns obtained from a spiral slot in a flat conducting sheet to those obtained from the sphere with two hemisphere cavities.

The radiation patterns from the spiral slot in the flat conducting sheet are bidirectional with essentially a complete null in the plane of the conducting sheet. When this spiral antenna is backed by a resonant cavity it becomes unidirectional with the main lobe still normal to the conducting sheet and a null in the plane of the conducting sheet.

When the spiral antenna is projected onto the hemisphere with the cavity produced by closing off the hemisphere at the equator, the radiation pattern is essentially unidirectional with null fill-in at the equator where a complete null was present in the case of the flat conducting sheet. There is a small back lobe caused by spill-over currents. When the two hemispheres are placed back to back and the proper phasing is accomplished as stated above, the side null is further filled in at the equator and the antenna becomes essentially omnidirectional, since the pattern is a figure of revolution about the polar axis.

At the higher frequencies, where the spacing of the center of radiation of each hemisphere is greater than 180°, a slight double null on each side is noticeable on account of the out-of-phase fields at some angle between the poles and the equator. These nulls are not generally very deep because of the large unbalance in contributing fields from each hemisphere.

A considerable amount of experimental testing was done to determine the total radiation field at four frequencies. Because the time scale was too short to make a detailed analysis of the theory and basic design, rather detailed experimental work was done. The efficiency of radiation was checked by comparison with a dipole radiating element and, as expected, the efficiency is a few db below that of a resonant dipole where the circumference of the sphere is less than one and one-half wavelength. However, for the frequencies, above the cutoff frequency, the gain was comparable with that of a one-half wave dipole.

The polarization of radiation from a double spiral slot antenna projected onto each hemisphere of a sphere is essentially circular from the poles and progresses through elliptical polarization to linear at the equator. The linear polarization from the equator is parallel to the polar axis. In order to produce the uniform phase distribution around the equator, the spiral on one hemisphere must be wound right hand and the other left hand as shown in Fig. 5(a). This results in right circular polarization from one hemisphere and left circular polarization from the other. This property makes the use of a circularly polarized receiving antenna impractical and, therefore, a linear polarized receiving antenna would probably yield the best results even with the cross-polarized null at the equator. Of course, polarization diversity would yield the best results. Figs. 7–10 show typical radiation patterns at each of the four frequencies for both horizontal (θ) and vertical (ϕ) polarization. Fig. 11 shows the coordinate system used in the radiation pattern measurements.

IMPEDANCE CHARACTERISTICS

The broad-band antenna also has fairly uniform impedance characteristics with frequency. Generally, the impedance will go through some sharp resonances at certain frequencies below the so-called cutoff frequency and then will exhibit fairly flat resistance and reactance above this frequency to the upper limit.

The cutoff frequency corresponds to the lowest fre-

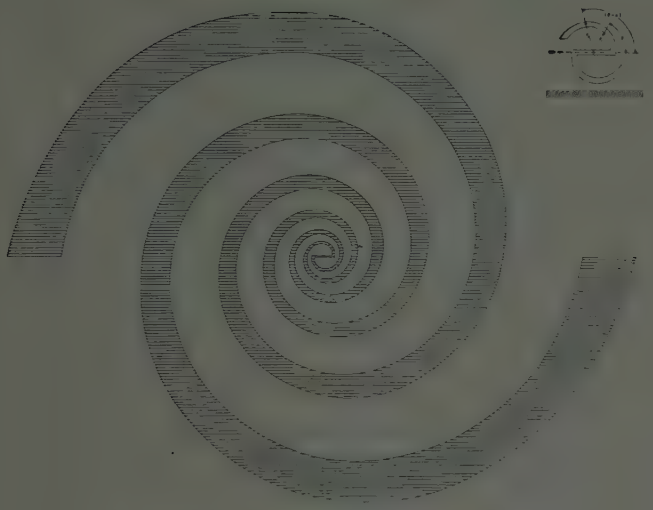


Fig. 3—Equiangular spiral slot antenna on a flat sheet.

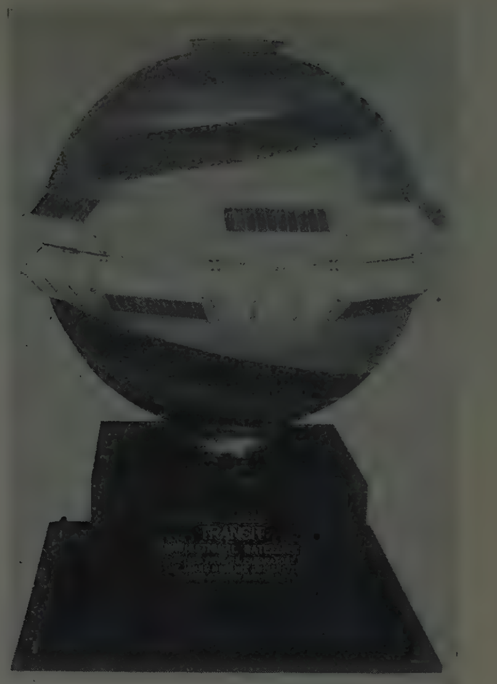
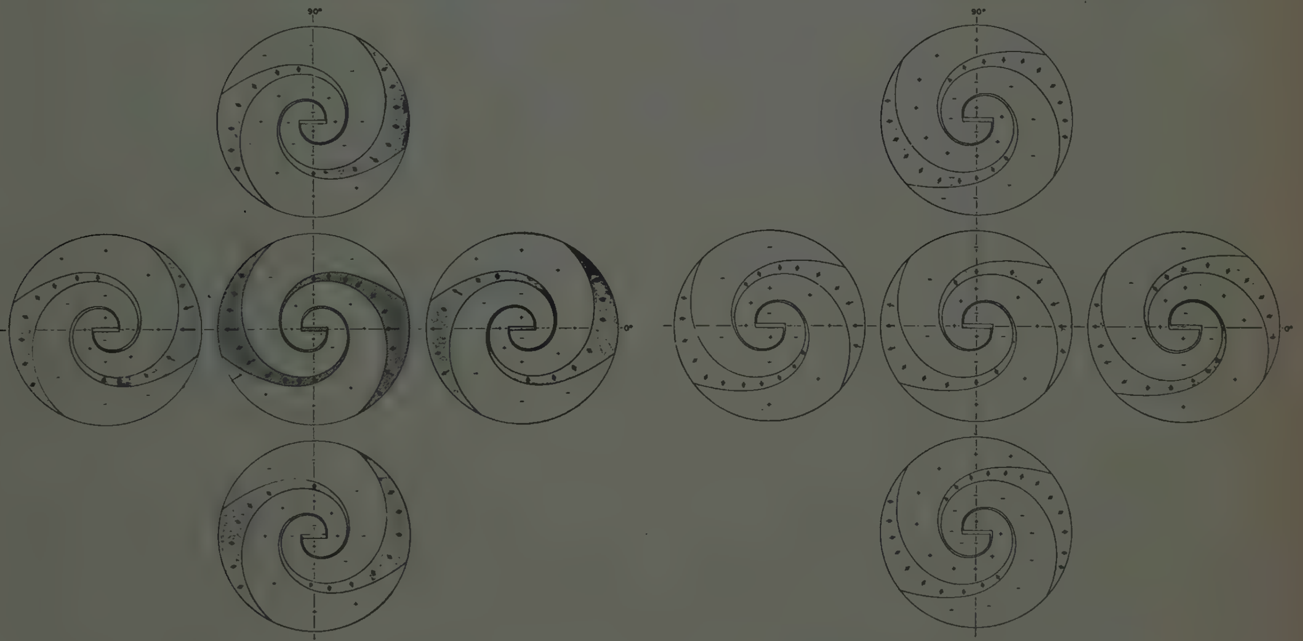


Fig. 4—TRANSIT satellite showing spiral slot antenna.



(a) (b)

Fig. 5—(a) Phasing diagrams of slot antenna clockwise and counter-clockwise spiral; (b) phasing diagrams of slot antenna two clockwise spirals.

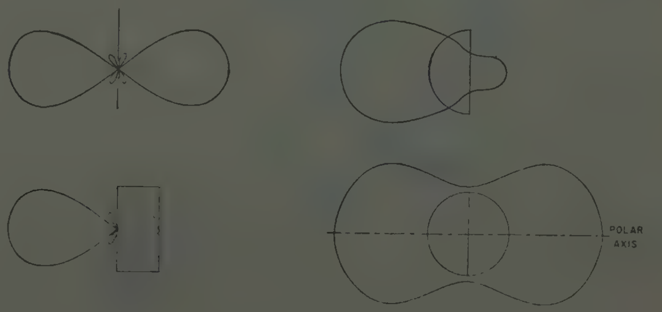


Fig. 6—Evolution of the spiral spherical antenna.

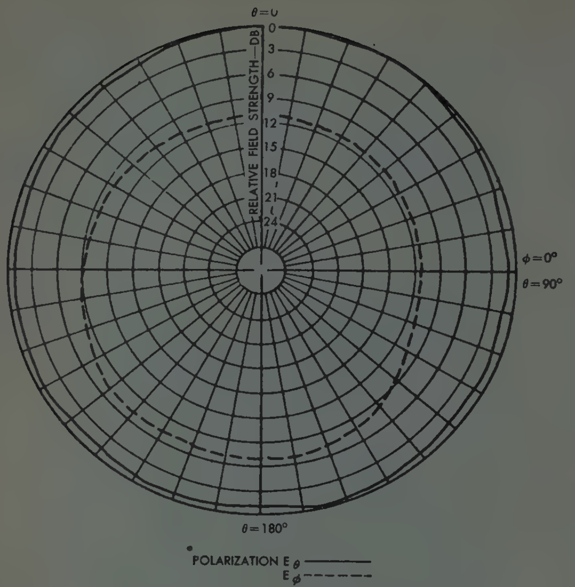


Fig. 7—Spiral antenna, frequency 54 mc.

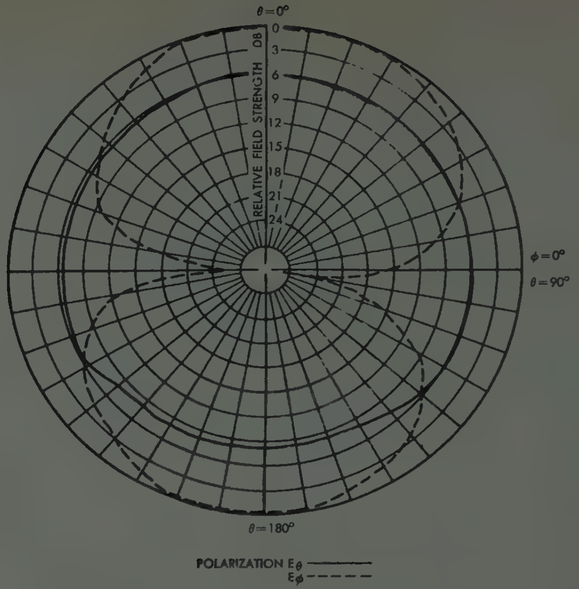


Fig. 8—Spiral antenna, frequency 108 mc.

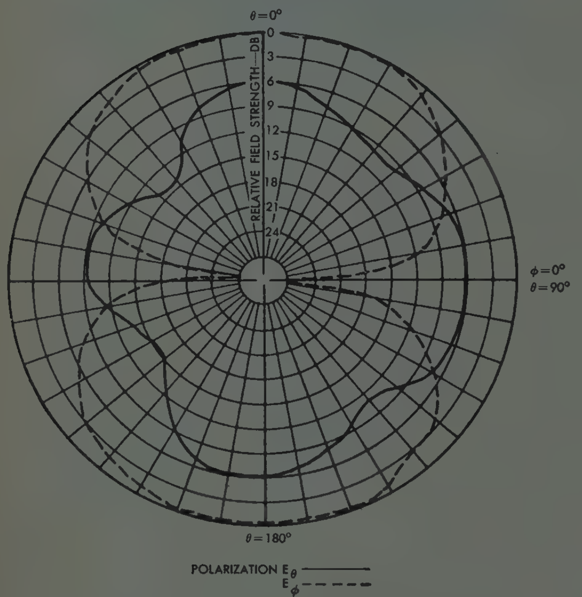


Fig. 9—Spiral antenna, frequency 162 mc.

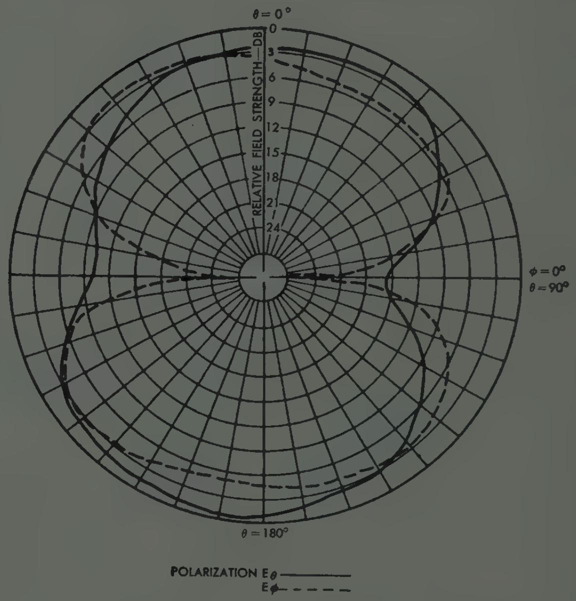


Fig. 10—Spiral antenna, frequency 216 mc.

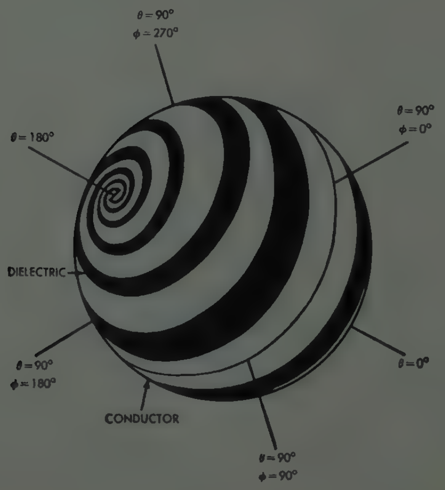

θ=90°
φ=270°
θ=180°
θ=90°
φ=0°
DIELECTRIC
θ=90°
φ=180°
θ=0°
CONDUCTOR
θ=90°
φ=90°

Fig. 11—Spherical coordinate radiation pattern reference system for satellite antenna.

quency at which the wave travelling down the slot is infinitely attenuated by the time it reaches the end. This is similar to the so-called infinite transmission line. As mentioned above, cutoff occurs at a frequency where the diameter of the spiral is approximately one-half wavelength or the circumference is about one and one-half wavelengths. For the 36-inch sphere used in the TRANSIT satellite, the cutoff frequency has a wavelength of about 2 meters or approximately 150 mc. The impedance measurements shown in Fig. 12 indicate that the cutoff frequency (that frequency at which the radiation resistance begins to flatten out) is about 135 mc.

The operation of this antenna at frequencies of 54 and 108 mc, therefore, is below the cutoff frequency, resulting in a high VSWR and lowered efficiency. However, after impedance matching, the efficiency has proven to be adequate as shown by field strength and gain measurements with respect to a one-half wave dipole.

Coupling and Matching Network

The RF feed cables are attached to the conductive coating between the slots and are carried in this manner to the feed points where the outer conductor of the coaxial line is attached to one side of the slot and the inner conductor to the other. This provides a method of feeding the balanced slot with an unbalanced feed line without the use of a balun. These RF feed cables are of equal length to provide in-phase coupling and are joined at a common *T* connector.

The use of the antenna at the four frequencies required isolation networks in the RF coupling system. After isolation, individual matching stubs are used to yield an input impedance of 50 ohms resistance.

The coupling and isolation networks for the four transmitters are constructed of coaxial cable using open quarter wave-stub techniques for notch filters. This type of filter provides approximately 30 db of isolation at each frequency. Fig. 13 shows a schematic diagram of such a coupling and isolation network. The insertion loss at each frequency is about 1.0 db. The use of filter stubs on each feed line gives some mismatch in addition to the normal VSWR caused by the antenna impedance's not being exactly equal to the characteristic impedance of the transmission line. This total mismatch is corrected to 50 ohms by the use of a separate matching stub at each frequency after the isolation filters. Thus, it affects only that frequency.

Conclusions

The antenna described in this paper has been flight tested and has given field strengths approximately equal to the measured values. The configuration used is not necessarily the most optimum and there are plans to continue the experimental work to confirm further the design for higher frequencies. No work has been done on variations of slot width for adjusting the impedance level. Radiation pattern measurements are being continued to determine the performance of other combinations of slot orientation in one hemisphere with respect to the other as well as using the same spiral screw sense on each hemisphere. It is felt that this antenna has some definite advantages for broad-band operation with a relatively uniform field distribution.

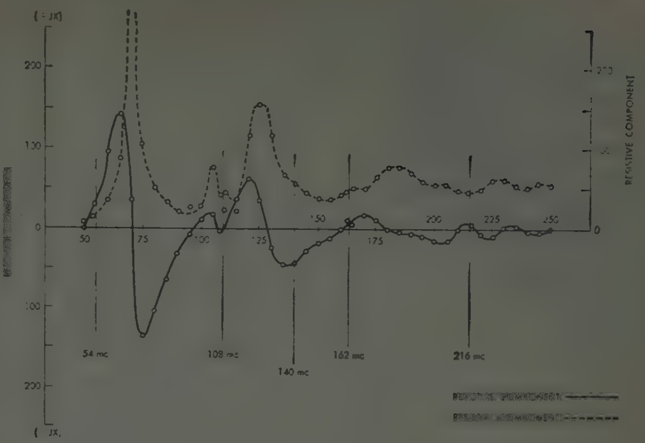


Fig. 12—Impedance data on APL/JHU sphere. Hemisphere with pole at $\theta=0°$ only.

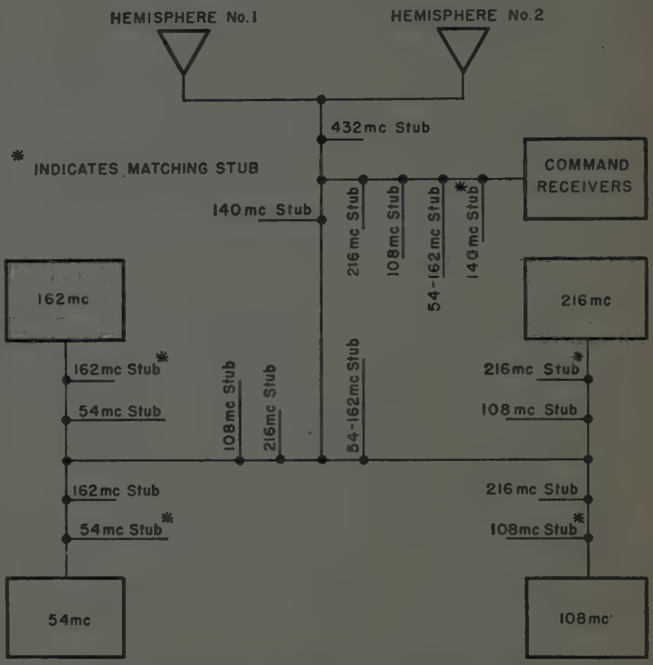


Fig. 13—Antenna matching and isolation network.

Acknowledgment

I wish to thank my associate in this work, R. M. Knight, Applied Physics Laboratory, The Johns Hopkins University, for his work and encouragement in performing many of the experimental tests, for performance evaluation. Also, the work of H. W. Haas and his associates at the Physical Science Laboratory, New Mexico State University, University Park, N. M., has been extremely helpful and is appreciated with many thanks.

Solar Batteries*

ARTHUR F. DANIEL†

***Summary*—The introductory part of this paper will provide a brief summary of how recent advances in military electronics for ground and space applications have created an urgent need for new energy sources and power supplies. In this search for new sources of power, considerable attention is being devoted to the utilization of solar energy.**

A short summary will be provided of the applications of USASRDL-developed solar cells in the satellite field to date. This will include a discussion of the apparent reliability and efficiency of solar cells in operating situations and the environmental factors affecting solar cells. A brief discussion will be given of the various methods of mounting solar cells with succinct comments on the advantages and disadvantages of different methods. Reference will also be made to USASRDL quality control methods and the quality vs cost problem.

The primary portion of the paper will be devoted to a detailed discussion of those energy conversion devices which employ static methods for direct conversion of solar energy to electrical energy, *i.e.*, those types which utilize the thermoelectric, Edison (thermionic emission), photovoltaic and photoemission effects.

The relative merits of these systems based on present and anticipated conversion efficiencies are reviewed, in brief, for guidance as to future possibilities of these types of devices and their utilization in the satellite program.

RECENT advances in military electronics for ground and space applications have created an urgent need for new energy sources and power supplies. In this search for new sources of power, considerable attention is being devoted to the utilization of solar energy.

Under the title "Solar Batteries," those energy conversion devices which employ static methods for direct conversion of solar energy to electrical energy will be discussed, *i.e.*, those types which utilize the photovoltaic, photoemission, the thermoelectric, and the Edison (thermionic emission) effects.

The relative merits of these systems based on present and anticipated conversion efficiencies will be reviewed with emphasis on space applications. The theory of these systems has already been well covered in a number of presentations by various authors and this information is readily available in the literature [1]–[13].

PHOTOVOLTAIC EFFECT

The photovoltaic effect has been successfully employed for the conversion of solar energy to electrical energy in Vanguard I, Sputnik III, and Explorers VI and VII, through the use of the silicon photovoltaic cell, quite often called a solar cell or solar battery, originally developed by the Bell Telephone Laboratories [14], [15]. A number of space vehicles to be launched in the near future will also depend on this type of solar power supply.

Silicon solar cells having efficiencies in the order of 9 to 10 per cent are now obtainable in reasonable quantities. In outer space, for earth orbiting satellites, solar energy is available at the rate of 2 cal/cm^2/minute or 1400 watts/meter^2 or 440 BTU/square foot^2/hour, with the receiver at normal incidence. Assuming an efficiency of 10 per cent (sea level reference 1.5 cal/cm^2/minute cell temperature at 25 degrees C) 14 mw/cm^2 of active cell surface is realized. In an actual design of a solar power supply, consideration is given to the type of orbit, duty cycle, spinning or oriented vehicle, cell temperature, erosion of protective covers by micrometeorites (dust), spectral shift from sea level to outer space, and protection for the cells during preflight tests and handling in addition to conditions existing during the launching period. After corrections for the above conditions, 8 mw/cm^2 or 7.4 watts/square foot of active cell area is generally considered available. A power-to-weight ratio of 0.25 watt/pound, including storage system, for a nonoriented satellite was achieved in an actual solar power supply, shown in Fig. 1 mounted on the satellite. The power supply [16], including the sealed nickel cadmium storage batteries, weighed $22\frac{1}{2}$ pounds. The nonoriented solar power supply was rated at 5 watts and weighed $17\frac{1}{2}$ pounds. If orientation were possible, using similar design methods, a power to weight ratio of 1.5 watts/pound would have been possible instead of the 0.25 watt/pound which was realized.

Looking into the near future, it is expected that 1×2-cm cells with efficiencies in the range of 12 to 15 per cent can be obtained in reasonable quantities. Because of the expected future higher power requirements, oriented expansible arrays of cells must be employed. The use of oriented arrays affords new possibilities in limiting the equilibrium temperature of the cells to the order of 0°C or lower. This low cell temperature can be achieved through the use of spectrally selective optical coatings which would increase the emissivity of the cell surface in the far infrared and reject radiant energy in the portions of the spectrum not utilized by the cell; it can be achieved by providing a good thermal path from the back of the cell to the rear surface of the array and by coating the rear of the array to obtain high surface emissivity to take advantage of the cooling effect of space. Even lower cell temperatures are a definite possibility. Initial information from the "paddlewheel" (Explorer VI) satellite indicates that the temperature of the cells is about 8°C.

The efficiency of a photovoltaic cell increases with a decrease in temperature. For instance, a cell with an efficiency of 12 per cent at 25°C will have an efficiency of 13 per cent at 0°C. Assuming that cells with an efficiency

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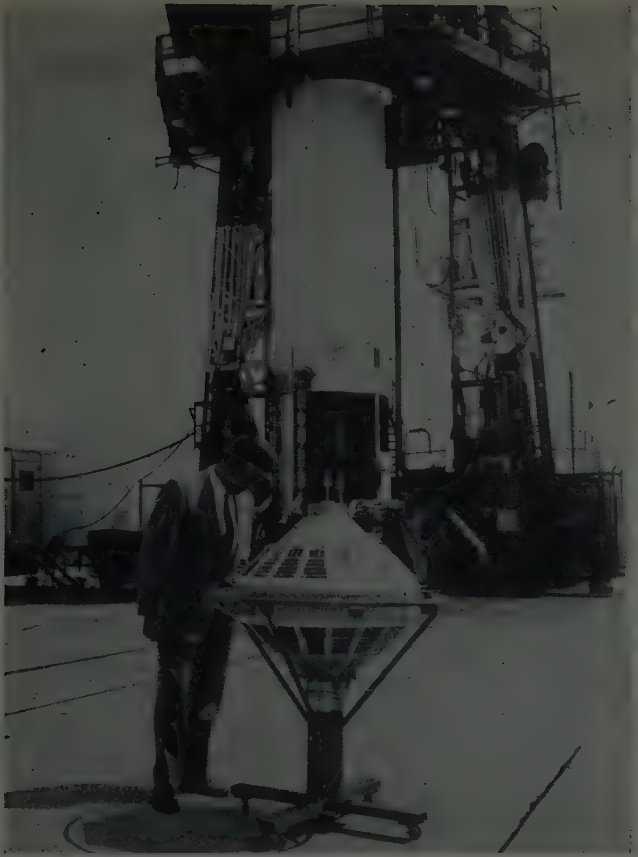

Fig. 1—Explorer VII showing solar power supply. Dr Kurt Debus, director of the Army Ballistic Missile Agency's Missile Firing Laboratory, looks over the Explorer VII satellite at the launching site. The 91.5-pound satellite was placed in orbit by the Juno II rocket shown in the background (U. S. Army photograph).

of 12 per cent (25°C) are available and that a cell temperature of 0°C can be obtained, an output of 18 mw/cm² of active cell surface can be realized. Compensating for transmission losses through cell covers sand blasted by meteoritic dust and the so-called spectral shift, the output is reduced to 14 mw/cm² or 13 watts/square foot. Using this figure with present design methods for the mounting of cells, orientation of plaque, and required protective covers and housing, a figure of 3.5 watts/pound (including storage batteries) can be calculated for the power output.

Programs are now in effect for the investigation of new materials for solar cells, multilayer and variable energy gap cells for greater utilization of the spectrum, high efficiency silicon cells, large area cells, multiple junctions on a single thin film of silicon, and high temperature cells. As a result of these efforts, efficiencies of 15 per cent for single gap cells and efficiencies of 20 to 30 per cent for cells of new materials and devices utilizing a greater portion of the spectrum may be possible. Present costs for photovoltaic cells are high and a program has been initiated for the investigation of methods leading to high production yields of efficient cells at a reasonable cost.

PHOTOEMISSION EFFECT

The photoemission effect [3], [4], [5], [6], [12] has been known for many years but little effort has been made in the past to use it for the conversion of radiant energy into electrical energy of sufficient magnitude for use as a source of power. A photoemission diode is analogous to the thermionic diode. When the photoemissive layer is illuminated, light photons release their energy to the electrons upon collision. If the amount of energy is greater than the work function of the material, the electron escapes with a kinetic energy equal to the excess of the quantum energy over the work function and it will reach the anode. In the thermionic diode, the anode is maintained at a lower temperature than the cathode, but in the photoemission diode, the anode is kept "dark," to prevent emission which would result in an opposing current. It is also subject to space charge limitations. Recently, several organizations have been investigating the possibility of increasing the efficiency of a device utilizing this effect, and 3 per cent appears obtainable. The chief advantages for this device are a high power-to-weight ratio and relatively low cost-per-unit power. Preliminary calculations indicate that approximately 3 kw of power at 28 volts can be obtained if the cells are mounted on a 30-foot diameter balloon. Fifty to 150 watts/pound appears feasible. The material costs are fairly low and large area devices may possibly be fabricated by vapor deposition of the photoemissive materials.

THERMOELECTRIC EFFECT

Solar powered thermoelectric generators are being considered as a source of power for space applications. Many groups are actively engaged in the investigation of thermoelectric materials and generator design. A survey of the literature and reports resulting from various programs sponsored by the Department of Defense indicates that a number of thermoelectric materials have been investigated and their efficiencies over a given operating temperature range have been determined. Fig. 2 and Fig. 3 show the efficiencies of N- and P-type materials in terms of a figure of merit, Q^2/pK, when $Q^2 = \text{volts} \times \text{deg}^{-1}$, $p = \text{ohm}^{-1} \times \text{cm}^{-1}$, $K = W \times \text{deg}^{-1} \times \text{cm}^{-1}$, which is based on the measured thermoelectric properties of the materials. It does not represent the over-all efficiency obtainable in a generator where losses caused by contact resistance between members of the thermocouples, lateral heat losses, and transfer of energy from the source to the hot junctions must be taken into consideration. For example, the over-all efficiency of a solar powered thermoelectric generator will be

$$\text{Efficiency} = \frac{\text{Electrical output}}{\text{Total intercepted energy}}.$$

It can be seen in Figs. 2 and 3 that the materials with the highest figures of merit are P-type $BiSbTe_3(Se)$ and

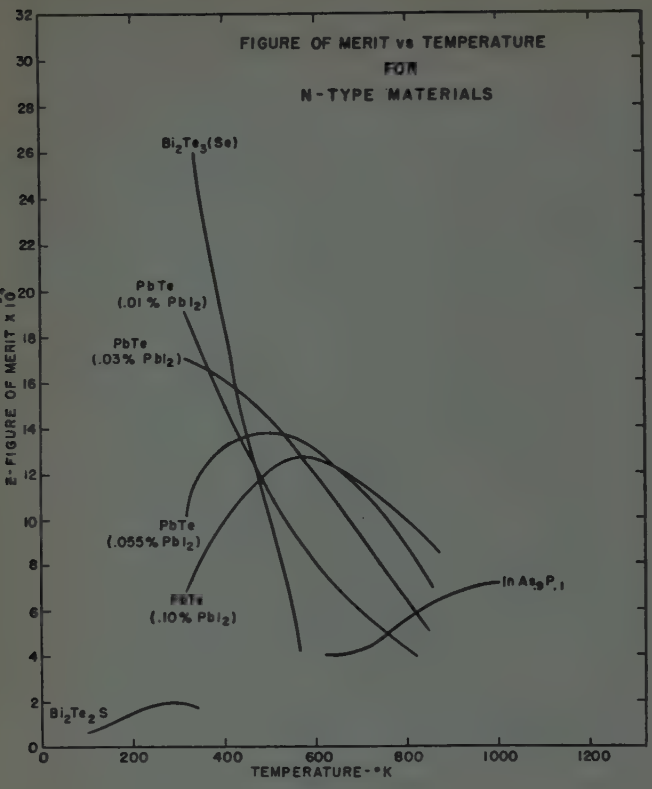

Fig. 2.

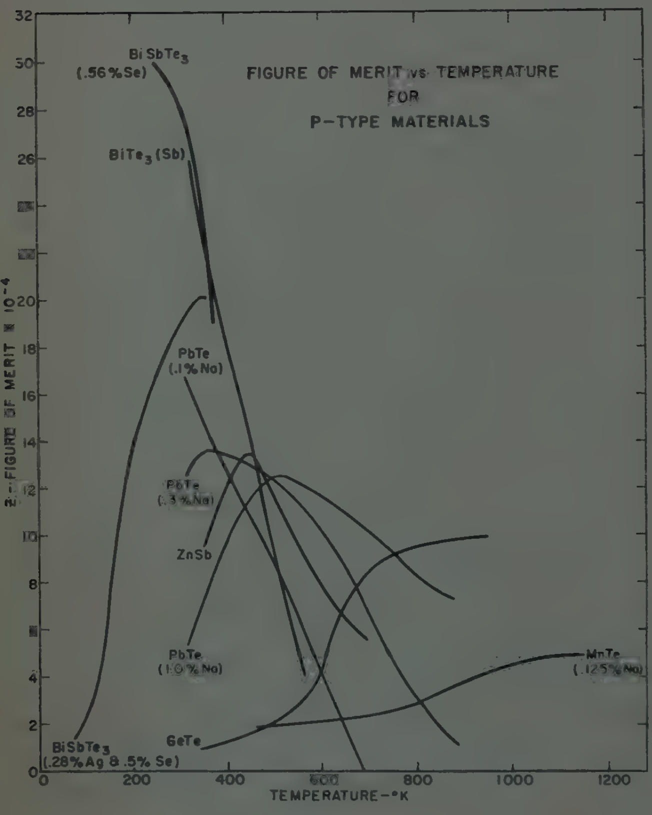

Fig. 3.

N-type Bi_2Te_3(Se). Because of the low melting points of these materials, however, their use in a solar-powered thermoelectric generator does not appear feasible. Materials capable of operating at higher hot junction temperatures are required in order to obtain a reasonable Carnot efficiency. A recent paper [17], discusses the merits and design problems of a solar-powered thermoelectric generator and its possibilities for space use. Since the data presented was obtained from the performance of an actual device, it will be used later in this discussion for comparison to other systems.

THERMIONIC EMISSION

Considerable effort is being devoted to the utilization of the Edison effect as a means of converting thermal energy into electrical energy. A number of laboratories are actively engaged in this field and claims of efficiencies of 10 per cent for the close spaced vacuum diode have been made and predictions of efficiencies up to 40 per cent for the Cesium vapor type have appeared. As for its present and future possibilities in space applications, a recent paper [3] gave a comprehensive engineering analysis of a thermionic power supply using solar energy as a heat source. The data presented will be used as a basis for comparison with other systems later in this discussion.

COMPARISON OF CONVERSION SYSTEMS

In comparing the relative merits of these various systems for the direct conversion of radiant energy from the sun to electrical energy, a number of factors must be taken into consideration. Some of these are listed below:

Reliability,
Power per unit weight, area or volume,
Complexity of the system,
Damage by meteorites; punctures and erosion,
Radiation damage,
Operating lifetime,
Mechanical problems introduced by use of expansible arrays, concentrators, pumps,
Accuracy of orientation,
Effects of drag and radiation pressure on large collector surfaces,
Effects of zero gravity,
Conversion of electrical output to usable form.

The basic difference between these four systems is that the photoelectric and photoemission devices convert the direct incident radiation. The thermoelectric and thermionic devices are essentially heat engines and require the concentration of the incident radiation to the device or heat sink to achieve the required rate of heat flux per unit of cathode or heat sink area. The use of solar concentrators introduces a number of problems that require serious consideration. For temperatures in the order of 500 to 750°C, required for thermoelectric generators, small errors in collector geometry and orientation can be tolerated. For temperatures in the order of 1000°C and above for thermionic diodes, a high degree of accuracy in collector geometry and orientation is essential for reasonable collection efficiencies. Thermal energy conversion systems must include a means for transferring the heat from the receiver or sink to the conversion device possibly by means of a circulating heat transfer fluid and transferring the waste heat to a radiator for rejection to space. Circulating pumps will be required to accomplish this in a zero gravity environment.

Consideration has been given to various types of energy storage systems capable of furnishing power during the dark periods of the orbit. The storage of thermal energy has been suggested as an alternate means for the electrochemical storage battery. If the power requirements for a particular vehicle are such that the load is a continuous one, a thermal energy storage system may be satisfactory. If peak power demands are involved, a thermal energy storage system will not be adequate because of the thermal lag in such a system and, in addition, the conversion device will have to be designed for this maximum power. This will introduce a voltage regulation problem, particularly for the thermoelectric generator. The output voltage would change considerably from peak load to average load. It appears more practical to use an electrochemical storage system which will be capable of supplying the peak loads with adequate voltage regulation and design the conversion device to supply the average power on the light side of the orbit with additional capacity for the recharging of the storage batteries.

Another problem involving the orientation of solar concentrators or solar cell arrays is the degree of accuracy required. For a solar cell array, a deviation of 20° from normal incidence only results in a loss in power of approximately 6 per cent. For solar collectors with fairly high concentration ratios, accuracy of orientation of 1° or less is necessary. This imposes a severe requirement on the orientation system and indicates a need for accurate resetting on the dark side of the orbit in order to minimize the time required for orienting the solar collector upon entering the light side of the orbit. If this time is appreciable, it may represent a significant portion of the light side period of the orbit.

The effects of erosion or punctures by micrometeorites present problems for all four systems. The erosion of the active surfaces of photovoltaic cells can be prevented by providing them with protective, transparent, and radiation resistant covers. Allowances must be made for the transmission losses in this cover, including the losses that will eventually occur when it becomes frosted because of sandblasting. The effects of punctures resulting in a loss of a cell or cells by larger particles can be compensated for. The cells are usually arranged in series-parallel groups and allowances can be made in the design for loss of a certain number of modules. The possibility of erosion by dust and sputtering caused by the impact of high energy atoms or molecules of the highly reflective surfaces of a solar concentrator must be taken into consideration. The possibility of providing a protective cover for solar concentrators does not appear feasible. Punctures would remove only a very small fraction of the total collector surface area. In applications where heat transfer loops are used from the source to the conversion device and from the device to a heat exchanger for rejection of waste heat, punctures of these components by meteorites would tend to be catastrophic, particularly in those using a liquid heat transfer medium.

Radiation damage to the various components employed in these systems does not appear to be a serious factor. The solar cells on Vanguard I have been operating continuously since March 17, 1958. The Russians [2] claim that the solar cells on Sputnik III have not been affected by radiation. The use of organic materials in construction of these systems introduces a radiation damage problem. This problem is difficult to solve because there is insufficient knowledge of the deleterious effects that may occur in a large portion of the ultraviolet spectrum. A high vacuum environment may also cause a degradation of organic materials. The use of spectrally selective coatings for the heat sink to make it an efficient absorber or as a means to obtain a low ratio of absorbtivity vs emissivity for solar cells and the structural materials of the array is being investigated. The stability of these coatings under radiation and a high vacuum environment and under conditions of high humidity prior to launching will have to be determined.

The four solar energy conversion systems being considered here do not carry their prime energy source with them as is done in chemical or nuclear energy conversion systems. Rather, the solar energy is converted to electrical energy at the point at which it is available. The essential factors to consider then would be the power per unit of weight and per unit of area, system complexity and reliability, operating lifetime of the conversion device, effects of space environment and cost. The one system that has been proven in actual use in space applications is that employing the photovoltaic cell and electrochemical storage batteries [16]. Information for use in comparing the merits of the other systems to that of the photovoltaic cell is in the form of engineering estimates based on present results obtained on experimental models and predicted future capabilities of the conversion device. With this in mind, some approximate figures, as shown in Table I, can be arrived at to indicate the possibilities of these four systems.

At the present time, the photovoltaic system appears to offer the highest watts/square foot and highest system efficiency and the lowest watts/pound. The thermionic and thermoelectric systems require solar concentrators and the efficiency of these devices depends on the

TABLE I

System	Watts/Pound		Watts/Square Foot of Collector Intercept Area		Possible System[4] Efficiencies	Life of Conversion Device
	Present	Future	Present	Future		
Photovoltaic	1.5	5	7.4	13.5	10–12 per cent	2 years or more
Thermionic [1]	7.2[1]	17	2.2	3.5	2– 4 per cent	Not determined
Thermoelectric [17]	20[1,2,3]	30[2,3]	0.3	1	2– 3 per cent	Not determined
Photoemission	—	50–150[3]	—	3	2– 3 per cent	Not determined

[1] Experimental.

[2] The watts/pound are only for the thermoelectric materials.

[3] Does not include storage system.

[4] System efficiency is defined as $\text{Efficiency} = \frac{\text{Electrical output}}{\text{Total intercepted energy}}$.

accuracy of the reflector surfaces and orientation. Efficiencies of 60 per cent have been indicated for concentrators assuming geometrical perfection and proper alignment. Efficiencies in the order of 30 to 40 per cent appear more likely for large expansible concentrators. This low concentrator efficiency imposes a penalty on the thermionic and thermoelectric systems. This is indicated in Table I where the power output in terms of watts/square foot of collector intercept area is shown for the four systems. This is further illustrated by comparing the required collector intercept area for a thermionic system [1] where 4680 square feet are required for a 10-kw power supply employing diodes with a 5 per cent efficiency. If photovoltaic cells covered an equivalent area and produced 7.4 watts/square foot, the power output would be in the order of 34 kw. To produce 10 kw, only 1350 square feet would be required. To be competitive with the other systems, more efficient, high temperature materials will be required for the thermoelectric system. The photoemission system is in its early stages of development. Present indications are that it could offer the highest watts/pound and possibly eliminate the need for orientation if mounted on a balloon, but would require large surface areas.

Conclusion and Recommendations

It is not possible, at this time, to reach any decision as to which of the direct energy conversion systems will eventually be superior. A number of factors have to be considered. System efficiency is only important where it eventually determines total system weight. If weight is no longer of major importance and the problems of storing and protecting large collectors during launch, their erection and orientation in space, and the effects of radiation pressure and drag in low altitude orbits, are major ones, then system efficiency becomes an important factor. Reliability and lifetime of the conversion device are of major importance. The photovoltaic and photoemission systems appear to have advantage over the other more complex systems. The effects of radiation and damage by micrometeorites are common problems for all systems. A zero gravity and high vacuum environment will be a greater problem for systems employing heat transfer loops and radiators for rejection of waste heat. It does not seem reasonable to assume that any one system will be used for all space applications.

Although all of the solar energy conversion systems are in relatively early stages of their development, the photovoltaic system has frequently been referred to as the "Work Horse" of these systems, and it may have to carry the major load of direct solar energy conversion for a number of years. Because of the state of development, an accurate and realistic estimate of the future practical possibilities of these systems is impossible at this time. Research and development in these areas should be vigorously pursued in order that the most effective system can be determined and the best solar energy conversion devices can be developed for space applications.

Acknowledgment

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Radiative Cooling of Satellite-Borne Electronic Components*

JAMES R. JENNESS, JR.†

Summary—**The basic principles of radiative cooling of electronic components and subassemblies in a satellite are discussed, and estimates are made of the lowest temperatures attainable in a satellite by completely passive means. It appears feasible to maintain some compartments within a satellite at temperatures of 250°K or lower, so an opportunity is presented for refrigerating components whose characteristics are enhanced at lower temperatures.**

INTRODUCTION

NOISE in many electronic components is reduced at lower temperatures. In view of the current attention being directed toward electronic equipment for satellite installation, it is of interest to estimate the lowest temperatures which can be obtained by completely passive means in a satellite.

SATELLITE HEAT BALANCE

In a satellite at stable-orbit altitudes the only sources of heat are internally-dissipated power, solar radiation, earth radiation and perhaps lunar radiation. (Stellar radiation is insignificant.) A previous study[1] has shown that if a satellite has an outer skin coated with a material (such as white porcelain enamel) having a high reflectivity in the spectral region of most intense solar radiation and high emissivity at longer infrared wavelengths, its skin temperature will become no higher than 55°C at the central "hot spot" on the side toward the sun. A spherical satellite with such a skin coating, which distributes absorbed heat uniformly around the skin by conduction or by rotation to give uniform exposure to solar radiation, will be at a temperature of about −40°C. Hass, Drummeter, and Schach[2] have found that the skin temperature might drop as much as 40°C while passing through the earth's shadow. Therefore, −60°C (213°K) might be taken as the average skin temperature, and the lowest temperature at which a non-heat-dissipating electronic component insulated from the skin can be maintained. However, these estimates of the temperature neglect the effect of internally-dissipated power. If the combined effect of all additional heat sources is equal to that of solar radiation, the minimum temperature at which a non-heat-dissipating component can be maintained is of the order of $2^{1/4}(213°\text{K}) = 253°\text{K}$ or −20°C. The actual average temperature probably is lower than this, but higher than −60°C.

The above temperatures apply to small satellites which receive fairly uniform exposure to solar and earth radiation. The attainment of temperatures below those estimated here will require a larger satellite whose orbit and attitude minimize the exposure of a certain part of the skin to both solar and earth radiation.

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* Original manuscript received by the IRE, December 2, 1959.

† HRB-Singer, Inc., State College, Pa.

[1] J. R. Jenness, Jr., "The effect of surface coating on the solar radiation equilibrium skin temperature of an earth satellite," *Solar Energy*, vol. 2, pp. 17–20; July/October, 1958.

[2] G. Hass, L. F. Drummeter, Jr., and M. Schach, *J. Opt. Soc. Amer.*, vol. 49, pp. 918–924; September, 1959.

Assuming that an isolated section of the skin is not exposed to solar or earth radiation, let us evaluate the effect of lunar radiation. It is evident that at a location near the earth the maximum intensity of solar radiation reflected by the moon, plus solar radiation absorbed by the moon and re-emitted, can be no greater than $SA/2\pi R^2$, where S is the solar constant, A is the area of the lunar disk, and R is the distance from the earth to the moon. Inserting the values of A and R, it is found that the maximum intensity of lunar radiation is 1.11 (10^{-5}) of the solar constant. A gray body exposed to radiation of this intensity and no other heat source will stabilize at 22.6°K. If this radiation falls on a surface at 50°K, it will raise the surface temperature to $(22.6^4+50^4)^{1/4}$°K, which is less than 1°K above the initial temperature of 50°K. Therefore, we can conclude that the moon need not be considered as a heat source if the refrigerated components are to be at a temperature no lower than 50°K.

Some Radiative Cooling System Design Considerations

If the orbit and attitude of the satellite are such that neither solar nor earth radiation falls on the radiative heat sink, the low temperature limit will be imposed by thermal conduction from warmer parts of the satellite. Consider the cooling system schematic shown in Fig. 1. It will be assumed that the thermal gradients in the heat-conduction path are small so that the cooled component or subassembly, conducting bar, and radiator are all at the same temperature T_c. Then the equation governing the temperature is

$$k \frac{A_s}{b}(T_o - T_c) = A_r \epsilon \sigma T_c^4$$

where k is the thermal conductivity of the insulating material of thickness b, A_s is the total area of the insulated surface separating the cooling system from the rest of the satellite's volume, T_o is the temperature at the outside of the insulating layer, A_r is the area of the external radiator, ϵ is the emissivity of the radiator, and σ is the Stefan-Boltzmann constant.

If the insulating layer is of styrofoam, $k=0.25$ BTU hour^{-1} in feet^{-2} °F^{-1}. Reasonable values of A_s and b are 3 feet^2 and 2 inches. By internal radiation exchange, T_o will be approximately the same as the average skin temperature, 253°K. It will be assumed that $T_c=153$°K and $\epsilon=0.9$. Since $\sigma=1.825$ (10^{-8}) BTU foot^{-2} °K^{-4} hour^{-1}, it is found that A_r must be 7.5 feet^2.

With the values of the parameters chosen in the above example, radiative cooling may be difficult. Without investigating how low the thermal resistance of the cooling system must be to distribute the heat uniformly over so large a radiator, it is evident that such a value of A_r implies a larger value of A_s than was used in the example, unless the radiator is mounted outside the satellite. To keep such a radiator out of the sunlight, severe restrictions must be imposed on the satellite's attitude and orbit or a cumbersome assembly of external shielding baffles must be mounted around it.

The required value of A_r can be decreased by thicker insulation. Thus, if b is increased by a factor of 2 or 3 in the example, A_r becomes approximately the same as A_s, reducing the radiator to manageable proportions. In practice it may be necessary to pot the entire system in foam as shown in Fig. 2, with the outer skin in separate sections to reduce conduction of heat around the skin from warmer areas.

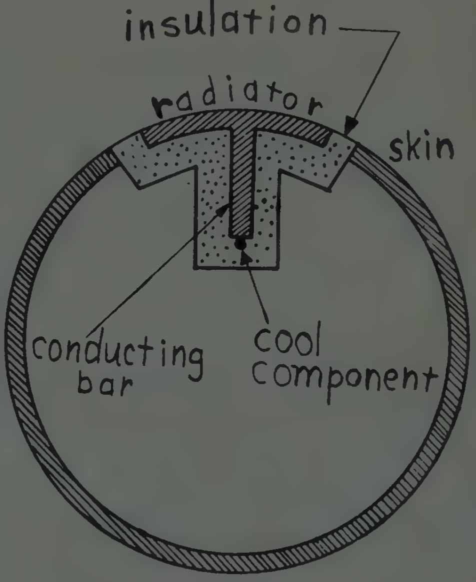

Fig. 1—Radiative cooling system schematic.

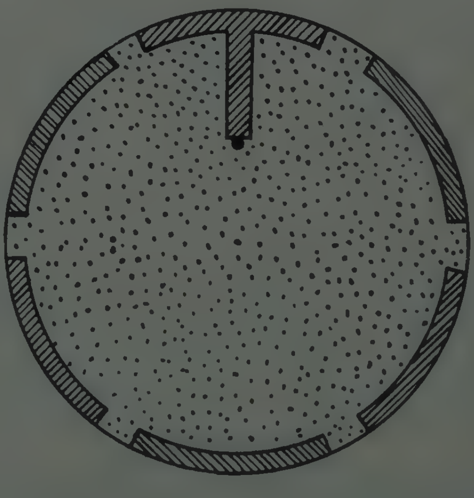

Fig. 2—Radiative cooling system potted in insulating foam in satellite with segmented skin.

CONCLUSION

With the proper skin coating, the attainment of temperatures below 250°K in a small satellite appears feasible if the ratio of internally-dissipated power to skin area is not excessive. In such a small "cool" satellite, components which require higher temperatures for proper operation must be in good thermal contact with heat-dissipating components, with relatively high thermal resistance between them and the skin.

If a larger satellite's orbit and attitude are such that an isolated section of its skin receives no radiation from the earth or sun, it might be possible to refrigerate a few components passively to temperatures as low as 150°K, keeping heat-dissipating components (and components which require higher temperatures) in contact with warmer skin sections. However, in addition to the obvious difficulties inherent in the attainment of such an orientation of the satellite, there will be problems in achieving structural strength while maintaining high thermal resistance between the refrigerated compartment and the remainder of the satellite's volume. Strengthening members will constitute thermal leaks through the insulation. Also, low thermal resistance in the conducting bar and radiator are to a certain degree incompatible with light weight.

Nevertheless, it may be feasible to maintain some electronic components and subassemblies in a satellite at temperatures below 250°K, so the possibility of taking advantage of these lower temperatures should be considered by designers of electronic equipment.

Extra-Terrestrial Radio Tracking and Communication*

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R. L. CHOATE†, MEMBER, IRE, AND L. R. MALLING†, MEMBER, IRE

***Summary*—When the U. S. Army lunar program was initiated in 1958, the Jet Propulsion Laboratory (JPL) was assigned the responsibility for the upper rocket stages and the payload. Payload responsibility included radio tracking and communication with the lunar probe.**

JPL's Microlock system, used for communicating with the early Explorer satellites, did not have sufficient range to perform this mission. Therefore, a new radio system designated TRAC(E) (TRacking And Communication, Extra-terrestrial) was designed. To the best knowledge of the authors, the TRAC(E) system is the first deep-space communication link to provide accurate tracking and telemetry data at lunar distances.

The design principles of the TRAC(E) system are presented in this paper in conjunction with a description of the equipment and actual performance data taken during the Pioneer IV lunar mission in March, 1959. The TRAC(E) system is an integral part of the NASA/JPL radio tracking station located near Goldstone Lake north of Barstow, Calif.

Future plans for improving the performance of the TRAC(E) system are indicated.

I. INTRODUCTION

THE TRAC(E) system is a versatile space communications system which is capable of simultaneous tracking and communication with earth satellites, lunar vehicles, or space probes. Two major questions which had to be answered in the design of such a communication system for extreme range applications were: What is the operating frequency? What is the size of the ground antenna? The ground antenna is considered in a separate report.[1] The factors which determined the operating frequency of the TRAC(E) system are outlined herein.

The transmitter in the space probe produces a signal power (P_t) limited by the efficiency of the transmitter and the weight of the vehicle electrical power supply. This power is beamed in the direction of the Earth-based receiver to provide an effective gain (G_t) of the signal power relative to radiation uniformly into space. The amount of beaming that can be realized is limited by the accuracy with which the vehicle's antenna can

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† Jet Propulsion Lab., California Institute of Technology, Pasadena, Calif.

[1] K. W. Linnes, W. D. Merrick, and R. Stevens, "Ground antennas for space communication systems," IRE TRANS. ON SPACE ELECTRONICS AND TELEMETRY, vol. SET-6, pp. 45–54; March, 1960.

be pointed and/or its attitude controlled. Since the feasibility of vehicle-borne directional antennas for space communications has not yet been demonstrated, present vehicle antennas are gain-limited. Therefore, the signal power (P_r) received by an Earth-based station is given by

$$P_r = P_t G_t A_r \left(\frac{1}{4\pi R^2} \right). \tag{1}$$

The term R represents the distance between the space vehicle and the Earth-based station. The Earth-based antenna is assumed to be area-limited.[1] Examination of (1) reveals that the received power is independent of frequency, to a first order at least.

The choice of the operating frequency can be narrowed somewhat by choosing frequencies which are not seriously affected by the ionosphere (above 100 mc) nor by the atmosphere (below about 10,000 mc). With regard to interference, virtually every known source of radio interference decreases with increasing frequency. Of particular importance to space communications is the decrease in brightness temperature of the galaxy with frequency. The largest source of galactic interference lies along the Milky Way or galactic center in the band approximately 15 degrees wide lying near the possible destination and orbits of space vehicles. Reduction in this interference can be accomplished by using antennas with relatively narrow beamwidths and operating at the highest frequency that is practical.

In view of the factors considered above, the optimum communication frequency was actually determined by such factors as vehicle transmitter efficiency and ground receiver sensitivity, that is, state-of-the-art considerations. Within the time scale of the U. S. Army Lunar Program (which was subsequently transferred to NASA) and the weight limitations of the Pioneers III and IV Lunar probes, development of an efficient stable transmitter in the 1000-mc range appeared feasible. An operating frequency of 960.05 mc was eventually chosen.

The design of the TRAC(E) system is based upon phase-lock techniques which have been under continuing investigation and development at the Jet Propulsion Laboratory (JPL) since 1953.[2–5] The technique, used by JPL in the Microlock system,[6] permits use of receiver bandwidths significantly smaller than the Doppler shift expected from space vehicles and proper addition of a limiter provides near-optimum performance over a wide dynamic range of both signal and interference levels.[3] Action of the VCO is such that with proper design, the Doppler component is effectively removed at the first mixer which permits narrow-band predetection amplification. Telemetry subcarriers are synchronously detected, and the information is available at the output of the phase-locked FM discriminators.

II. TRAC(E) System for Pioneers III and IV

Fig. 1 is a functional block diagram showing the TRAC(E) system as it was originally designed for tracking and communicating with the Pioneers III and IV probes. In the probe transmitter, the 40.0021-mc crystal-controlled oscillator signal is phase modulated with telemetry subcarriers which contain the measured data as frequency modulation. The resultant phase-modulated signal is frequency multiplied by a factor of 24 and amplified to provide a 960.05-mc signal to the probe antenna. The transmitter signal output, normalized to unity, is

$$\mathrm{Cos}\ [24\omega_c t + s(t)] \tag{2}$$

where $s(t)$ is the composite phase modulation signal.

The TRAC(E) receiver is a narrow-band phase-coherent double-conversion superheterodyne receiver which employs three balanced mixers at 960 mc, three balanced detectors at 30 mc, and four synchronous detectors at 455 kc. The mixers and detectors are three terminal devices having an input, a coherent reference and an output, and they perform the analytic function of multiplying the waveform of the input signal by the waveform of the reference. These devices in conjunction with three 30-mc IF preamplifiers, three 455-kc IF amplifiers, and the voltage-controlled local oscillator comprise the major portion of the ground receiving system. They are arranged in such a manner as to function as the differencing elements of four highly accurate servo systems: 1) the RF servoloop at 960 mc, 2) the AGC loop, 3) the declination servomechanism, and 4) the hour angle servomechanism.

The received signal is

$$A(t) \cos\ [24(\omega_c t - \phi_c) + s(t)], \tag{3}$$

where $A(t)$ is the signal amplitude factor of the space-to-earth signal at the ground station and

$$\phi_c = \frac{\omega_c}{c} \int_0^t \dot{R} dt,$$

radians, the Doppler shift on ω_c. The significant waveform equations for the TRAC(E) receiving system are listed in Table I. The numeral preceding each expression refers to the corresponding numerals in Fig. 1.

The system parameters which were used to achieve

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[2] E. Rechtin, "The Design of Optimum Linear Systems," Jet Propulsion Lab., California Institute of Technology, Pasadena, External Publication No. 204; April, 1953.

[3] R. Jaffe and E. Rechtin, "Design and performance of phase-lock circuits capable of near optimum performance over a wide range of input signal and noise levels," IRE Trans. on Information Theory, vol. IT-1, pp. 66–76; March, 1955.

[4] C. E. Gilchriest, "Application of the phase-locked loop to telemetry as a discriminator or tracking filter," IRE Trans. on Telemetry and Remote Control, vol. TRC-4, pp. 20–35; June, 1958.

[5] S. G. Margolis, "The response of a phase-locked loop to a sinusoid plus noise," IRE Trans. on Information Theory, vol. IT-3, pp. 136–144; March, 1957.

[6] H. L. Richter, R. Stevens, and W. F. Sampson, "A Minimum Weight Instrumentation System for a Satellite," Jet Propulsion Laboratory, California Institute of Technology, Pasadena, External Publication No. 376; May 23, 1957.

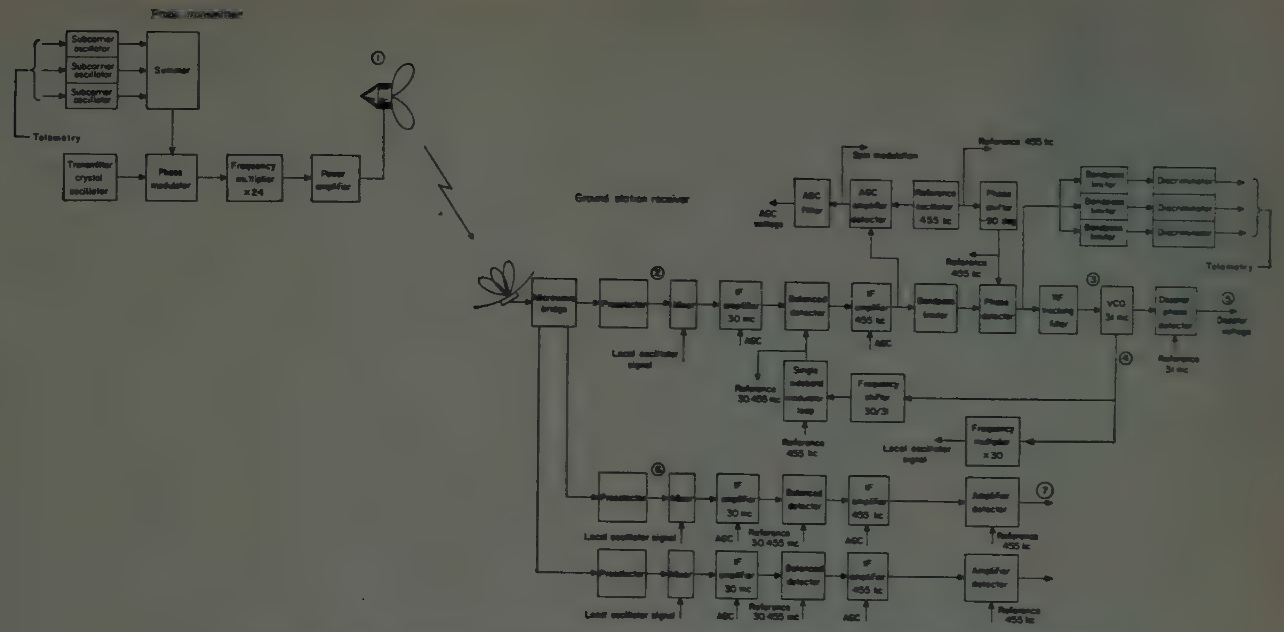
Probe transmitter
Subcarrier oscillator
Subcarrier oscillator
Subcarrier oscillator
Summer
Telemetry
Transmitter crystal oscillator
Phase modulator
Frequency multiplier ×24
Power amplifier
①
Ground station receiver
Microwave bridge
Preselector
②
Mixer
IF amplifier 30 mc
AGC
Local oscillator signal
Balanced detector
IF amplifier 455 kc
AGC
Bandpass limiter
Phase detector
IF tracking filter
③
VCO 31 mc
Doppler phase detector
⑤
Doppler voltage
④
Reference 31 mc
AGC filter
AGC voltage
AGC amplifier detector
Spin modulation
Reference oscillator 455 kc
Reference 455 kc
Phase shifter 90 deg
Reference 455 kc
Bandpass limiter
Discriminator
Bandpass limiter
Discriminator
Bandpass limiter
Discriminator
Telemetry
Single sideband modulator loop
Reference 30.455 mc
Reference 455 kc
Frequency shifter 30/31
Frequency multiplier ×30
Local oscillator signal
⑥
Preselector
Mixer
IF amplifier 30 mc
Balanced detector
IF amplifier 455 kc
Amplifier detector
⑦
Local oscillator signal
AGC
Reference 30.455 mc
AGC
Reference 455 kc
Preselector
Mixer
IF amplifier 30 mc
Balanced detector
IF amplifier 455 kc
Amplifier detector
Local oscillator signal
AGC
Reference 30.455 mc
AGC
Reference 455 kc

Fig. 1—Functional block diagram of the Pioneer IV transmitter and the Goldstone TRAC(E) System.

TABLE I
WAVEFORM EQUATIONS USED IN TRAC(E) SYSTEM

$$A(t)\cos\left[24\omega_c t+s(t)\right] \tag{1}$$

$$g_\Sigma(\alpha)A(t)\cos\left[24(\omega_c t-\phi_c)+s(t)\right] \tag{2}$$

$$\sin 24(\phi_c^*-\phi_c)\approx 24(\phi_c^*-\phi_c) \tag{3}$$

$$\cos\frac{31\times 3}{30\times 4}(\omega_c t-\phi_c^*) \tag{4}$$

$$\cos\frac{31\times 3}{30\times 4}\phi_c^*=\cos\left[\frac{31\times 3}{30\times 4}\right]\frac{\omega_c}{c}\int_0^t \dot{R}dt \tag{5}$$

$$g_\Delta(\alpha)A(t)\cos\left[24(\omega_c t-\phi_c)+s(t)\right] \tag{6}$$

$$E\frac{A(t)}{A^*(t)}\frac{g_\Delta(\alpha)}{g_\Sigma(\alpha)}\approx\frac{A(t)}{A^*(t)}g_\Delta'(0) \tag{7}$$

$A(t)$ = signal amplitude factor of the signal radiated from the probe

E = dc reference in the receiver AGC loop

c = velocity of propagation, meters/second

f_1 = measured frequency of the 31-mc Doppler reference

f_{c_0} = probe transmitter oscillator frequency with zero Doppler

f_d = measured frequency at the output of the Doppler phase detector

$g_\Sigma(\alpha)$ = voltage pattern of the main-beam antenna normalized so that $g_\Sigma(0)=1$

$g_\Delta(\alpha)$ = voltage pattern of the split-beam antenna normalized with respect to $g_\Sigma(0)$

$g_\Delta'(0)$ = slope of the $g_\Delta(\alpha)$ curve near $\alpha=0$, mils^{-1}

$s(t)$ = telemetry phase modulation

t = time, seconds

$\dot{R}$ = radial component of probe velocity meters/second $=c\left[1-\frac{960}{31f_{c_0}}(f_d+f_1)\right]$

α = tracking error of the antenna servo, mils

ϕ_c = Doppler phase shift on $\omega_c=\frac{\omega_c}{c}\int_0^t \dot{R}dt$, radians

ω_c = angular frequency of the probe transmitter oscillator $=2\pi(40.0021$ mc) radian/second

$\frac{1}{A^*(t)}$ = a measure of the voltage gain of the ground receiver so defined that the incoming signal $A(t)$ multiplied by the receiver gain figure $1/A^*(t)$ equals unity

accurate tracking and communication with the Pioneer IV probe to lunar distances and farther are listed in Table II. The 180 mw of transmitter power (±0.5-db power tolerance) was distributed between the carrier (96 mw) and the three telemetry carriers so that accurate tracking and reliable data transmission could be realized to at least 250,000 miles (lunar distance). Tolerances shown on the carrier (±0.3 db) and telemetry subcarriers (±0.5 db) apply to permissible variations in phase modulation index (±2° on each subcarrier). Examination of Table II shows that with the lunar probe at a range of 250,000 miles, the received carrier signal level is −141.1 dbm (decibels referred to one milliwatt). The resultant signal-to-noise ratios (S/N) in the RF loop, AGC loop, and angle tracking loop are 8, 39.7, and 38.2 db, respectively. This provided sufficient margin in signal level to insure successful tracking and communication. The significance of the bandwidths of the RF loop, AGC loop, and angle tracking loop shown in Table II is discussed later.

TABLE II

TRAC(E) SYSTEM CHARACTERISTICS FOR PIONEERS III AND IV

Transmitter Power, Total (at 250,000 miles)	180 mw±20 mw(+22.5 dbm±0.5 db)
Vehicle Antenna Gain	2.5 db±1 db
Carrier Power	96 mw±7 mw(+19.8 dbm±0.3 db)
Subcarrier Power, Channel No. 1	14 mw±1.5 mw(+11.5 dbm±0.5 db)
Subcarrier Power, Channel No. 2	14 mw±1.5 mw(+11.5 dbm±0.5 db)
Subcarrier Power, Channel No. 3	36 mw±4 mw(+15.5 dbm±0.5 db)
Space Loss at 250,000 Miles	204.5 db
Ground Antenna Gain	41.8 db
Transmission Line Loss	0.7 db
Receiver Noise Figure	7.5 db
Receiver Bandwidth ($2B_{L_0}$) at Receiver Threshold	20 cps
Receiver Threshold	−154.1 dbm
Angle Track Loop BW ($2B_L$)	0.06 cps
Angle Track Loop Receiver Threshold	−179.3 dbm
AGC Loop BW ($2B_{L_0}$) at Receiver Threshold	0.025 cps
AGC Loop Threshold	−183.1 dbm
At 250,000 Miles	
Received RF Signal Level	−141.1 dbm
S/N for RF Loop	8.0 db
S/N for AGC Loop	39.7 db
S/N for Angle Tracking Loop	38.2 db

III. DESIGN OF TRACKING RECEIVER

The specifications for the TRAC(E) tracking receiver for Pioneers III and IV are shown in the Appendix. Reference is made to the literature[2,5,7–10] with regard to the development of these specifications. The 20-cps RF phase-locked loop noise bandwidth at threshold is determined by the stability of the transmitter and receiver oscillators. As a consequence, the sensitivity of the receiving system is limited by the threshold of the RF phase-locked loop. With an effective noise temperature (T_{eff}) of 1435°K resulting from a receiver noise figure of 7.5 db, an apparent antenna noise temperature of 42°K and an 0.7-db loss for the simultaneous lobing RF bridge and antenna transmission line to the receiver input, the threshold of the receiver is −154.1 dbm. This assumes that the antenna is looking at a low-noise region of the sky.

The AGC system has a closed-loop noise bandwidth of 0.025 cps at receiver threshold. With this noise bandwidth, spin rate and spin modulation depth for the lunar probe are measured as an error signal in the AGC loop. In addition, rms gain error is maintained to 0.1 db at receiver threshold.

Gain and phase tracking in the angle tracking receiver relative to the reference receiver is ±2 db and ±10°, respectively, for input signal levels from −45 dbm to threshold. Gain tracking maintains near-constant gain in the hour angle and declination servosystems. Phase tracking maintains shift in the electrical axis of the antenna system (due to phase shift) to ±0.01 angular mils (±0.0006°) for a 1° antenna error pattern *s*-curve with a 40-db null depth. The 0.06-cps closed-loop noise bandwidth for the angle tracking servo-system shown in Table II maintains rms angle jitter to 0.4 angular mils (0.022°) at receiver threshold.

IV. DESCRIPTION OF EQUIPMENT

Probe Transmitter

The transmitter for the Pioneer IV probe was located in the central 6-inch diameter well of the instrumented payload (see Fig. 2). Fig. 3 shows the flight transmitter mounted on the webbed platform that forms part of the payload structure and also supports the shutter experiment. Transistorized subassemblies are mounted on micarta disks, interposed between the webbing to provide thermal isolation. The vacuum-tube cavity is fastened directly to the center web so that the whole payload becomes a heat sink for the vacuum tube. The overall weight of the platform and transmitter is 1.33 pounds. The transmitter less platform weighs 0.98 pound.

Circuit design was concentrated principally in the areas that would insure early completion of a 960-mc transmitter having the objective specifications outlined in Table III. Some of the specialized components and their features that helped make an efficient 960-mc lunar-probe transmitter possible are: 1) the Western Electric GF 45021 transistor, with its ability to operate efficiently at VHF frequencies (100 to 300 mc) previously restricted to subminiature tubes; 2) the GE-7077, a subminiature ceramic UHF triode, which permitted the design of efficient, subminiature UHF cavity amplifiers; 3) the Bell Telephone Laboratories' "Varactor," a silicon junction reactance diode developed for the U. S. Army Research and Development Laboratories, that operates as a frequency multiplier with considerably higher conversion efficiency than conventional diodes;

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[7] W. K. Victor, "The Evaluation of Phase-Stable Oscillators for Coherent Communication Systems," Jet Propulsion Lab., California Institute of Technology, Pasadena, External Publication No. 337; May 8, 1956.

[8] A. J. Viterbi, "Design Techniques for Space Television," Jet Propulsion Lab., California Institute of Technology, Pasadena, External Publication No. 623; April 13, 1959.

[9] W. B. Davenport, Jr., "Signal-to-noise ratios in band-pass limiters," *J. Appl. Phys.*, vol. 24, pp. 720–727; June, 1953.

[10] W. K. Victor and M. H. Brockman, "The application of linear servo theory to the design of AGC loops," PROC. IRE, vol. 48, pp. 234–238; February, 1960.

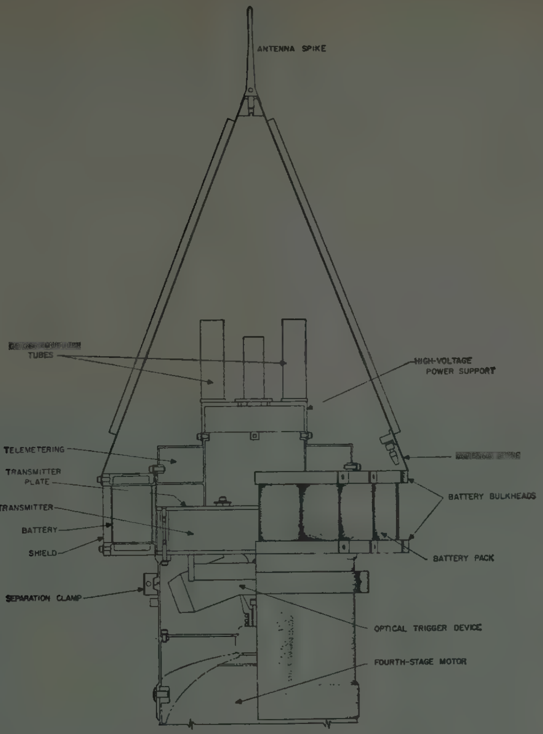

Fig. 2—Sketch of the Pioneer IV instrumented payload.

Fig. 3—Probe transmitter mounted on the webbed platform.

TABLE III

Probe Transmitter Objective Specifications

Carrier frequency (nominal)	960 mc
Output power	180 mw
Crystal oscillator frequency (nominal)	40 mc
Short-term frequency stability—15 minutes	1 part 10^7
Long-term frequency stability	1 part 10^6
Power output stability	$\pm\frac{1}{2}$ db
Maximum incidental phase modulation	2 degrees per g
Over-all weight	0.9 pound
Over-all size 6-inches diameter × 1-inch height	28 inches^3
Primary power input	2.4 watts
Operating time with 6.5-pound battery	120 hours
Transmitter efficiency, exclusive of antenna gain	7.5 per cent

4) the ruggedized quartz crystal developed by the U. S. Army Signal Engineering Laboratories that enables phase-stable signals to be generated under conditions of extreme vibration; and 5) the development of reliable voltage-sensitive capacitors which simplify the design of the phase-modulation circuits.

Semiconductor devices operating in circuits from 40 to 960 mc provided the RF signal drive to the vacuum-tube power-amplifier which provided RF excitation to the probe antenna. The objective specifications of the probe transmitter (Table III) were achieved in the Pioneers III and IV experiments. The probe transmitter is described in detail in the literature.[11]

Ground Tracking Receiver (*Goldstone Station*)

The RF portion of the TRAC(E) receiving system is mounted in an electronics enclosure on the rear portion of the antenna reflector structure[1] (see Fig. 4). The UHF portions of the reference, hour angle and declination angle receiver channels which include three preselector cavities, three balanced mixers and three 30-mc IF amplifiers, is housed in a weatherproof enclosure (see Fig. 5). The UHF portion of the local oscillator system which includes a ×6 frequency multiplier, a three-way power divider, three UHF attenuators for local oscillator drive adjustment, and a crystal current meter circuit is mounted in a second weatherproof enclosure. Other weatherproof enclosures (see Fig. 5) contain power supplies and test equipment. The test equipment, which includes UHF and VHF signal generators and a noise-figure test setup, facilitates periodic checks of the receiver RF circuitry to insure optimum performance.

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[11] L. R. Malling, "A 960-MC Lunar Transmitter," Jet Propulsion Laboratory, California Institute of Technology, Pasadena, External Publication No. 596; January 7, 1959.

The amplified 30-mc output signals from the reference, hour angle and declination angle channels are transmitted from the electronics enclosure through coaxial cables to the receiver rack in the control building. The receiver rack comprises three cabinets containing the remainder of the receiving system, power supplies, and associated instrumentation (Fig. 6). The receiver system can be operated with a closed-loop RF bandwidth of either 20 or 60 cps which can be selected by means of a switch. Another switch permits selection of either an 11-second or a 300-second time constant filter for the AGC circuit. The system was operated in the 20-cps bandwidth and 300-second AGC filter time constant mode for the Pioneer IV experiment (Table II).

Operational control of the tracking system during a tracking operation is provided at the control console (Fig. 7). Control of both the tracking receiver and antenna servosystem is provided at the console. In general, the receiver and servosystems each have three basic modes of operation—manual, automatic track, and acquisition. In the acquisition mode, the tracking receiver and antenna search in frequency and angle, respectively, about the predicted frequency and angular position. Following acquisition, the received signal is automatically tracked in frequency and angular position.[12] Various receiver monitoring functions are provided at the control console.

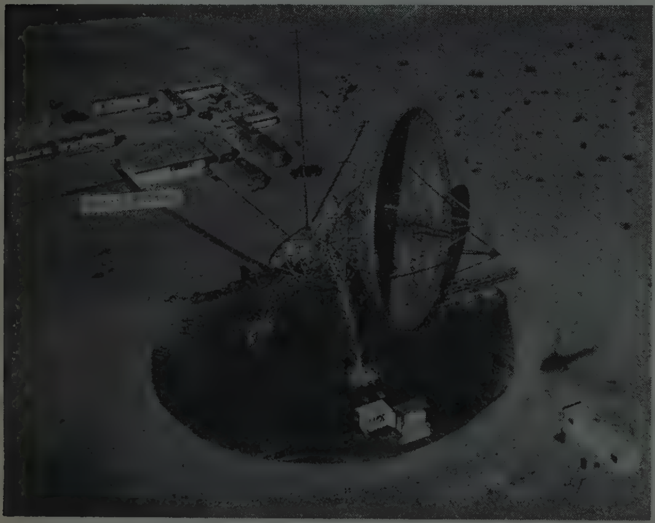
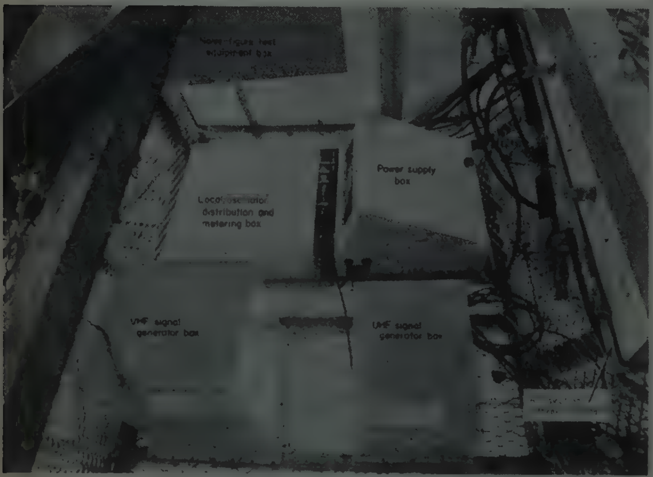


Fig. 4—Goldstone tracking station.


Noise-figure test equipment box
Power supply box
Local-oscillator distribution and metering box
UHF signal generator box
UHF signal generator box
Fig. 5—Electronics enclosure.

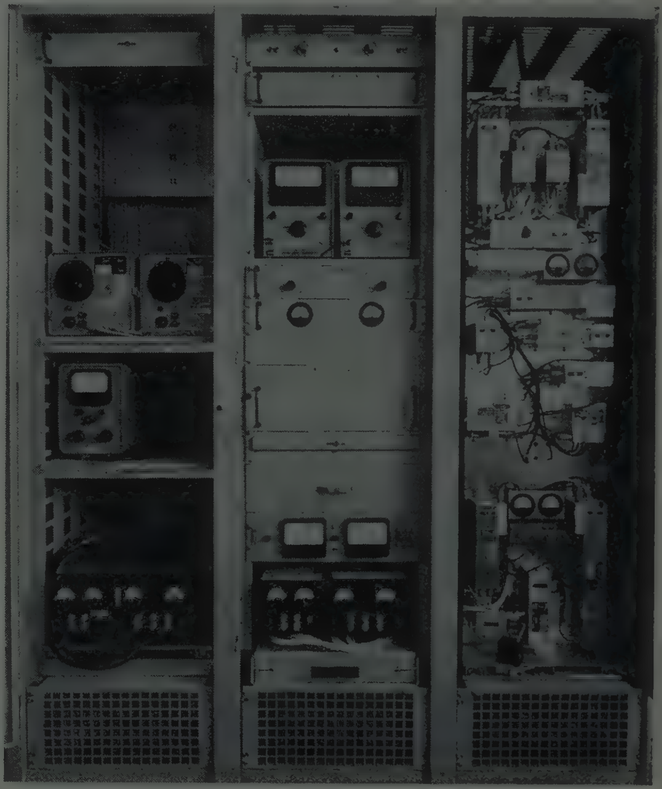


Fig. 6—Receiver rack.

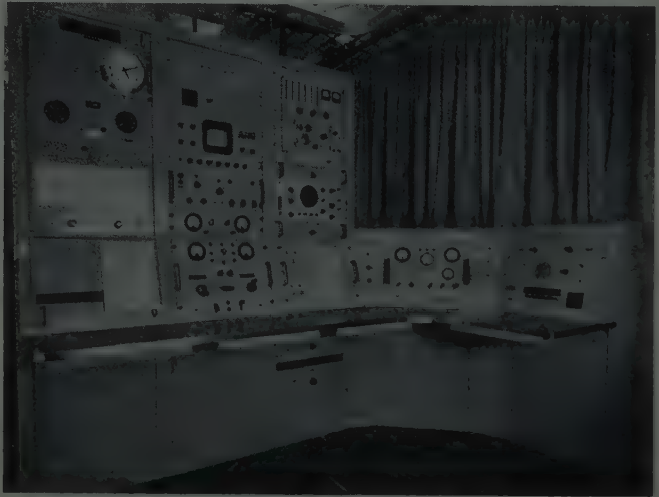


Fig. 7—Control console.

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[12] M. Eimer and R. Stevens, "Tracking and Data Handling for the Pioneer III and Pioneer IV Firings," Jet Propulsion Laboratory, California Institute of Technology, Pasadena, External Publication No. 701; August 14, 1959.

V. PERFORMANCE DATA FOR PIONEER IV LUNAR PROBE

Description of Pioneer IV Tracking

Following successful launching of Pioneer IV lunar probe at 05:10:56 GMT on March 3, 1959, radio tracking was accomplished through launch, injection, and coasting flight by the three JPL-operated stations (Cape Canaveral, Puerto Rico and Goldstone) out to a maximum range of 652,000 km (about 407,000 miles). At this range, after 82 hours and 4 minutes of flight, a rapid decrease in signal level occurred, as expected, due to loss of battery power. Prior to this event, received signal level was 5 db above threshold in a 20-cps bandwidth. After loss of signal during the fourth tracking period, a narrower bandwidth filter (10 cps) was installed in the tracking receiver, and the RF carrier was tracked for about 9 additional minutes. Had the batteries lasted, tracking could have been accomplished to a range of 1,850,000 km (1,150,000 miles) using the 10-cps filter.

Performance of the TRAC(E) tracking and communication equipment was highly satisfactory during this experiment. Performance data for the Goldstone Station are presented in the following material.

Received Signal Level and Spin Modulation: The noise figures of the Goldstone receiver based on data taken March 1 and 3, 1959, were as follows:

Reference Channel	7.5 db,
Hour Angle Channel	7.15 db,
Declination Channel	7.25 db.

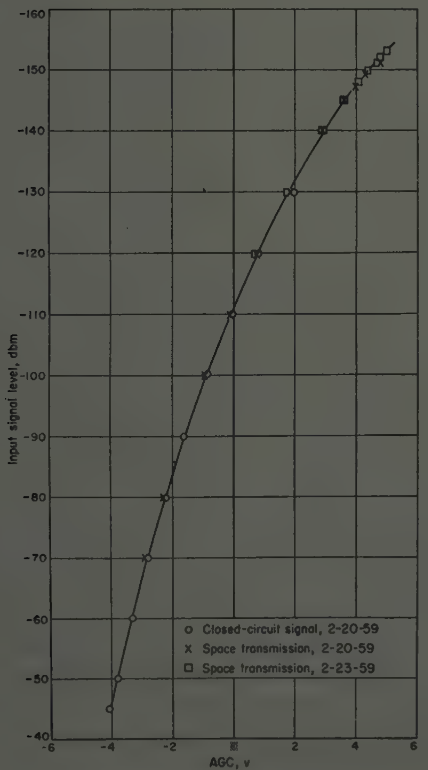
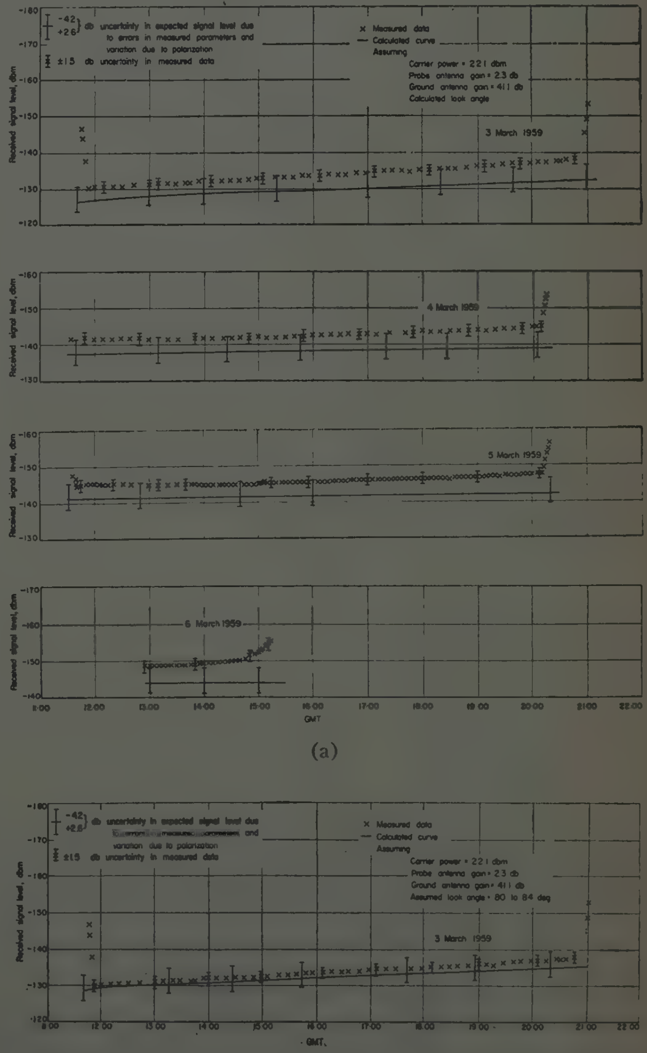


Fig. 8—Receiver sensitivity.

The measured noise figure of the reference channel agreed with the nominal value of 7.5 db. The theoretical receiving system sensitivity based on unity signal-to-noise ratio in a 20-cps bandwidth, a receiver noise figure of 7.5 db (effective noise temperature 1355°K), an apparent antenna noise temperature of 42°K and a total loss of 0.7 db for the antenna simultaneous lobing bridge system and transmission line to the receiver input is −154.1 dbm. The measured receiver sensitivity curve is shown in Fig. 8.

A comparison of measured with calculated signal level is shown in Fig. 9(a) for each tracking period for the Pioneer IV experiment. Throughout the tracking operation, the measured values are 3 to 5 db less than the calculated curves (based on nominal values and assumed vehicle attitude).

After despin at 16:32:24 GMT on March 3, the amplitude modulation of the signal resulting from spin and precession occurred at frequencies of 0.23 cps and 0.046



(a)

(b)

Fig. 9—Comparison of measured and calculated values of received signal level.

TABLE IV

RANGE AND SIGNAL LEVEL DATA FOR PIONEER IV

Tracking Period No.	Date	Time	Range Kilometers	Signal Level at Acquisition dbm	Signal Level at Loss of Lock dbm	Change of Signal Level (Neglecting Horizon Effect)	
						Measured db	Expected db
1	March 3, 1959	11:47:00 to 21:05:48	97,400 to 194,000	−144.5	−154.5	7.5	6.1
2	March 4, 1959	12:33:54 to 21:15:53	315,000 to 380,000	−141.5	−154.0	2.5	1.7
3	March 5, 1959	12:34:52 to 21:18:46	487,000 to 545,000	−147.5	−156.5	2.0	1.0
4	March 6, 1959	12:50:00 to 15:16:19	644,000 to 657,000	−150.5	−156.2	0.7	0.2

cps with a maximum modulation depth of 3 to 3.5 db and a minimum depth of 1 to 1.5 db. If the look angle between the spin axis and the line-of-sight to the probe is assumed to be 80° with a precession angle of 10°, the antenna pattern measurements agree with the observed depth of spin modulation, and the average signal strength comparison would appear as shown in Fig. 9(b) for the first tracking period.

The received signal levels were determined by measuring the AGC voltage with a digital voltmeter and converting this to signal level, using the receiver sensitivity curve (see Fig. 8). The measured values are indicated by the symbols in Fig. 9. In general, the slopes of the measured curves are slightly greater than the expected slopes. The signal levels at acquisition and loss of signal and the changes in level for each period are shown in Table IV. The estimated uncertainty of ±1.5 db in the measured data is based upon the following:

Receiver sensitivity error	±1 db,
Recording, calibration, and data reduction errors	±0.5 db.

In Fig. 9 the expected signal level is shown for comparison with the measured data. The solid line represents the signal level calculated on the basis of the following:

Transmitted carrier power	22.1 dbm
Vehicle antenna gain	function of look angle (see Fig. 10)
Ground antenna gain	41.1 db (indirect measurement)
Space attenuation (db)	$92.1+20\log_{10}$ (range in kilometers).

The look angle is calculated on the assumption that the spin axis of the probe is essentially colinear with the velocity vector at injection (look angle varies from 45.5 to 56.2° for the experiment). The vehicle antenna pattern of Fig. 10 is based upon the measurements of antenna gain as a function of look angle and rotation of the antenna. The solid curve which was used for calculating signal level represents the geometric mean of the gain for full rotation at the given look angle. The dashed curves show the maximum and minimum gain and indicate the degree of amplitude modulation to be expected at each given look angle. The uncertainty of the calculated data is based upon the following:

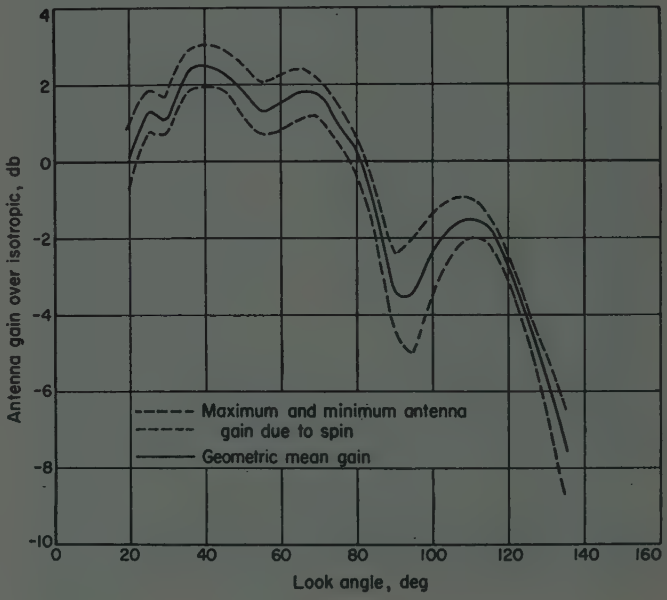


Fig. 10—Vehicle antenna pattern.

Ground antenna gain (variation due to polarization	$\left.\begin{matrix}-1.7\\+0.1\end{matrix}\right\}$ db
Ground antenna gain (uncertainty)	±1 db
Transmitter power	±0.5
Vehicle antenna gain	±1.0
This gives a total uncertainty in the calculated data of	$\left.\begin{matrix}-4.2\\+2.6\end{matrix}\right\}$ db.

The measured signal levels are less than the expected values by 3 to 5 db. While the regions of uncertainty of the calculated and observed data overlap slightly, it is believed that the difference between expected and measured values is of the order of 3 db. This apparent loss of signal can be explained by assuming a look angle of the order of 80°.

Angle Tracking System: The reduction of tracking data was essentially a problem of filtering, by statistical analysis, the random observational errors and the sys-

TABLE V
LONG-TERM SYSTEM CAPABILITIES

Characteristic	Capability		
	1958	1960	1962
Transmitter Power	0.2 watt	10 watts	100 watts
Vehicle Antenna Gain	0 db	16 db	36 db
Receiver Sensitivity:			
Noise temperature	2000° K	400° K	40° K
Bandwidth	30 cps	30 cps	30 cps
Earth-vehicle range for 10-db S/N ratio	4×10^5 miles	4×10^7 miles	4×10^9 miles

tematic bias errors. The computational program for generating the Pioneer IV trajectory is described in detail in a JPL external publication.[12] The trajectory angles were corrected for atmosphere refraction and all known mechanical sources of error then compared with unsmoothed raw data for the last day of tracking (March 6, 1959). The hour angle errors for this period ranged from 0.035 to −0.11°. Near threshold, the hour angle errors were +0.02 to −0.1°. Using smoothed values of observed data, the hour angle error was about −0.03° at the system threshold. The declination errors for the same period ranged from +0.07° to −0.04°. Using smoothed values of observed data, the declination error was about 0.02° at threshold. As indicated in an earlier portion of this paper, calculated rms angle jitter at system threshold is 0.4 mil, which generally agrees with measured values.

VI. FUTURE PLANS FOR IMPROVEMENT

Extra-terrestrial radio communications has been demonstrated and practiced successfully by radio astronomers for many years. As a result of their pioneering effort, the job of communicating with rocket-launched space probes has been and will continue to be much less difficult. Utilization of their data, analyses, techniques and in some cases actual equipment designs made it possible to develop the TRAC(E) system in a very short period of time with sufficient long-term capabilities that it could communicate to the distance of the moon in 1958 and to the planets during the period 1960–1962. (See Table V.) A basic element of the TRAC(E) system, for example, is the 85-foot diameter antenna designed for radio astronomy research by the Blaw-Knox Company under an Associated Universities contract. With this antenna and the TRAC(E) receiver described in this report, it would be possible to communicate to the nearby planets. However, future plans are to extend the range (and/or the bandwidth) still further by the addition of low-noise amplifiers. An over-all noise temperature of 400°K is predicted for 1960 using a parametric amplifier, and a temperature of 40°K for 1962 using a maser.

The real significance of Table V is that the principal factors which are expected to extend the communicating range are related to the space vehicle. Over a period of four years, it is suggested that the vehicle transmitter power could be increased 27 db, from 200 mw to 100 watts, and the vehicle antenna gain could be increased 36 db. With these improvements, Table V shows that space communications could be maintained anywhere in the solar system with a bandwidth of 30 cps. Solution of these problems in the time scale indicated depends on many factors, such as vehicle performance, payload weight, primary power, earth-seeking vehicle antenna system, financial support, etc.

The present TRAC(E) system is a one-way link which measures two angles and one-way Doppler, and receives telemetry. Future TRAC(E) communications will be two-way, employing a 10-kw UHF ground transmitter and vehicle-borne transponder. The two-way TRAC(E) system will measure two angles, range and range-rate; it will receive telemetry and also provide a radio command link to the space vehicle.

APPENDIX

TRAC(E) TRACKING RECEIVER DESIGN SPECIFICATIONS FOR PIONEERS III AND IV

I. Frequency—nominal 960.05 mc (also see XII).

II. Noise Figure—7.5 db or less.

III. Bandwidth—20 cps at threshold.

IV. Threshold—154.1 dbm defined as the signal level at which the receiver loses lock.

V. Range of Input Signal Level—−45 dbm to threshold.

VI. RF Loop Characteristics

A. RF Loop Transfer Function

$$H(s) = \frac{1 + \dfrac{3}{4B_L} s}{1 + \dfrac{3}{4B_L} s + \dfrac{9}{32 B_L^2} s^2}$$

where

$$2B_L = \frac{1}{2\pi} \int_{-\infty}^{\infty} | H(j\omega) |^2 dw$$

$$= \frac{2}{3} B_{L_0}\left(1 + 2 \frac{a}{a_0}\right)$$

and

$$2B_{L_0} = 2\left(\frac{9}{32} \frac{G_0}{\tau_1}\right)^{1/2}$$

$\tau_1 = R_1C$,

G_0 = open-loop gain at threshold
$= 360 \times 30 \times k_m \times k_{\text{vco}} \times \alpha_0$,

k_m = sensitivity of the RF phase detector in volts/degree
$= 0.7$ volt/degree,

k_{vco} = sensitivity of the RF VCO in cps/volt
$= 250$ cps/volt,

α_0 = signal suppression factor at threshold
$= 0.088$ for 20-cps bandwidth.

$$\alpha = \left[\frac{1}{1 + \text{antilog}_{10}\left(\dfrac{1 + P_N - P_s}{10}\right)}\right]^{1/2}$$

$$= \left[\frac{1}{1 + \text{antilog}_{10}\left(\dfrac{1 - 134.1 - P_s}{10}\right)}\right]^{1/2}$$

where

P_N = available noise power in the 455-kc IF amplifier 2-kc pass band (expressed in dbm)

k = Boltzman's Constant, 1.37×10^{-23} joule/°K
$= \log_{10}(kT_{\text{eff}}\Delta f \times 10^3) = -134.1$ dbm

P_s = available signal power in the same 2-kc bandwidth in the same units as P_N.

Consequently

$$H(s) = \frac{1 + \dfrac{1}{\lambda}s}{1 + \dfrac{1}{\lambda}s + \dfrac{1}{2\lambda^2}s^2}$$

where

$$\lambda = \frac{4}{9}B_{L_0}\left[1 + \frac{2}{\alpha_0}\,\frac{1}{1 + \text{antilog}_{10}\left(\dfrac{1 - 134.1 - P_s}{10}\right)}\right]^{1/2}.$$

B. RF Loop Noise Bandwidth

1) Design Value at Design Threshold and Signal Frequency ($2B_{L_0} = 20$ cps)

2) Variation with Signal Level

The RF loop noise bandwidth shall vary from 20 cps at design threshold to 158 cps at strong signal levels ($\alpha = 1$).

3) Variation with Signal Frequency

The threshold RF loop noise bandwidth shall vary from 20 cps at design signal frequency to not less than 14.1 cps at signal frequencies displaced ±2.6 parts in 10^5 from the design signal frequency.

C. RF Loop Phase Error

1) Variation with Signal Input Frequency

The static phase error at threshold shall not exceed 6° at signal frequencies displaced ±2.75 parts in 10^6 from the design signal frequency. At these two frequency extremes, the threshold will be reduced 1 db from that specified at design signal frequency. Static phase error at strong signal levels ($\alpha = 1$) shall not exceed 13.5° at signal frequencies displaced ±2.6 parts in 10^5 from the design signal frequency.

2) Maximum Rate of Change of Input Signal Frequency

The phase error at threshold shall not exceed 6° for a maximum rate of change of 6 parts in 10^9 per second in input signal frequency. At this maximum rate, the threshold will be reduced 1 db relative to that obtained for a zero rate of change in signal frequency.

The phase error at strong signals ($\alpha = 1$) shall not exceed 30° for a maximum rate of change of 3.5 parts in 10^7 per second in input signal frequency.

3) RMS Phase Error

The RMS phase error due to noise shall not exceed 1.0 radian at design threshold. The RMS phase error due to noise shall not exceed 0.06 radian at an input signal level of −120 dbm.

4) Residual Phase Error

The residual phase error shall not exceed 3° peak at strong signal levels ($\alpha = 1$) within an RF loop noise bandwidth of 20 cps.

5) Noise Bias

The maximum phase error due to noise at threshold shall not exceed 1 per cent of the phase detector S curve amplitude at threshold (equivalent to static phase error of 0.6°).

VII. AGC Loop Characteristics (Coherent AGC)

A. AGC Loop Transfer Function

$$H(s) = \frac{1}{\left(1 + \dfrac{1}{G}\right) + \dfrac{\tau}{G}s}$$

where

τ = time constant of AGC filter
$= 300$ seconds

G = gain of the AGC loop
$= K_DK_A$

and

K_D = AGC detector constant expressed in volts/db
$= 3.1$ volts/db at threshold
$= 4.0$ volts/db at −45 dbm input signal level

K_A = constant associated with the gain of the receiver expressed in db/volt
$= 5$ db/volt at threshold
$= 17$ db/volt at −45 dbm input signal level.

B. AGC Loop Noise Bandwidth

1) Design Value at Design Threshold

$2B_L = 0.025$ cps

2) Variation with Signal Level

The AGC loop noise bandwidth shall vary from 0.025 cps at design threshold to 0.11 cps at an input signal level of −45 dbm.

C. AGC Loop Gain Error

1) Static Gain Error

The variation in coherent detected receiver output shall vary not more than 6 db for the range of input signal levels from −45 dbm to threshold.

2) Maximum Rate of Change of Input Signal Level

The gain error at threshold shall not exceed 1 db for a maximum rate of change of 0.05 db per second in input signal level. The gain error at an input signal level of −45 dbm shall not exceed 1 db for a maximum rate of 0.25 db per second in input signal level.

3) RMS Gain Error

RMS gain error introduced in the RF and angle tracking loops at threshold shall not exceed 0.1 db rms. RMS gain error at an input signal level of −120 dbm shall not exceed 0.003 db rms.

VIII. 30/31-mc Frequency Shifter Phase-Locked Loop Characteristics

A. Frequency Shifter Loop Transfer Function

$$H(s) = \frac{1 + \dfrac{3}{4B_L} s}{1 + \dfrac{3}{4B_L} s + \dfrac{9}{32B_L^2} s^2}$$

where

$$2B_L = 2\left(\frac{9G_0}{32\tau_1}\right)^{1/2}$$

$\tau_1 = R_1 C$

$G = 360 \times 32 \times k_m \times k_{\text{vco}}$

$k_m =$ sensitivity of the frequency shifter phase detector in volts/degree

$= 0.25$ volt/degree

$k_{\text{vco}} =$ sensitivity of the frequency shifter VCO in cps/volt

$= 45$ cps/volt.

B. Frequency Shifter Loop Noise Bandwidth

1) Design Value at Design Frequency

$2B_L = 1340$ cps

2) Variation with Frequency

The frequency shifter loop noise bandwidth shall vary from 1340 cps at design frequency to not less than 950 cps at frequencies displaced ±3 parts in 10^5 from the design frequency.

C. Frequency Shifter Loop Phase Error

The static phase error shall not exceed 5° at frequencies displaced ±3 parts in 10^5 from the design frequency.

IX. 30.455-mc SSB Modulator Phase-Locked Loop Characteristics

A. SSB Modulator Loop Transfer Function

$$H(s) = \frac{1 + \dfrac{3}{4B_L} S}{1 + \dfrac{3}{4B_L} S + \dfrac{9}{32B_L^2} S^2}$$

where

$2B_L = 2(9/32\ G/\tau_1)^{1/2}$

$\tau_1 = R_1 C$

$G = 360 \times K_m \times K_{\text{vco}}$

$K_m =$ sensitivity of the SSB modulator phase detector in volts/degree

$= 0.7$ volt/degree

$K_{\text{vco}} =$ sensitivity of the SSB modulator VCO in cps/volt

$= 250$ cps/volt.

B. SSB Modulator Loop Noise Bandwidth

1) Design Value at Design Frequency

$2B_L = 1200$ cps

2) Variation with Frequency

The SSB modulator loop noise bandwidth shall vary from 1200 cps at design frequency to not less than 850 cps at frequencies displaced ±3 parts in 10^5 from the design frequency.

C. SSB Modulator Loop Phase Error

The static phase error shall not exceed 10° at frequencies displaced ±3 parts in 10^5 from the design frequency.

X. Angle Error Channel Characteristics

A. Gain Tracking

The differential gain between either angle error channel and the reference channel shall not exceed 2 db over the range of input signal levels from −45 dbm to threshold. Gain tracking is determined by comparison of the coherent output signal levels in the angle error channel to the reference channel output.

B. Phase Tracking

The differential phase between either angle error channel and the reference channel shall not exceed ±10° over the range of input signal levels from −45 dbm to threshold. Phase tracking is determined by comparison of the phase of the coherent output signal level in the angle error channel to reference channel phase output.

C. RMS Angle Tracking Error

RMS angle tracking error in the 0.06 cps noise bandwidth of the angle tracking servosystem shall not exceed 0.022 rms at threshold.

XI. Modulation Characteristics

A. Amplitude Response and Phase Symmetry of 2-kc Pass Band

The amplitude characteristic of the 455-kc IF amplifier shall be flat to ±1.5 db within the 2-kc pass band. The phase characteristic of the 455-kc IF amplifier shall be symmetrical to within 45° about the center of the 2-kc pass band.

XII. Frequency—The nominal center frequency shall be 960.05 mc with the ability to track plus or minus 3.0 parts in 10^5 from the nominal center frequency. By changing the crystal and retuning the VCO, the center frequency can be changed between 955 mc and 965 mc.

XIII. Input VSWR—The input VSWR to the receiver shall be less than 1.5 to 1 for a 50-ohm transmission line.

XIV. Image Rejection—The image rejection of the receiver shall be at least 40 db.

XV. Telemetry Bandwidth—The narrowest bandwidth of the receiver before the telemetry output shall be 10 kc.

XVI. Residual Phase Modulation—The residual phase modulation in the RF loop when measured with a signal source at -60 dbm shall be less than 3° peak.

XVII. Internal Interference in the RF Loop—The internal interference in the RF loop due to coherent leakage signals of the receiver shall be at least 40 db below the signal level at threshold when the interference enters ahead of the limiter and at least 60 db below the signal level at threshold when the interference enters following the limiter.

XVIII. Internal Interference in the AGC Loop—The internal interference in the AGC loop due to coherent leakage signals of the receiver shall be at least 40 db below the signal level at threshold.

XIX. Internal Interference in the Angle Detector Channels—The internal interference in the angle detector channels due to coherent leakage signals of the receiver shall be at least 40 db below the signal level at threshold.

XX. Cross Coupling Between Channels—The signal isolation between any two channels shall be at least 60 db.

XXI. Receiver IF Frequencies—The receiver IF frequencies shall be 30 mc and 455 kc.

XXII. First Local Oscillator Frequency—The first local oscillator signal shall be derived from a 31-mc VCO which is frequency multiplied by 30.

XXIII. Coherent Local Oscillator Frequencies—The local oscillator signals shall be derived in such a manner that the received signal frequency can be determined by measurement of the 31-mc VCO frequency only.

XXIV. Recording Outputs—Provision shall be made for making the following available for recording:

receiver AGC,
RF loop dynamic phase error and telemetry,
telemetry subcarriers,
Doppler frequency plus constant frequency,
spin modulation,
declination error,
hour angle error.

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Tracking and Display of Earth Satellites*

F. F. SLACK† AND A. A. SANDBERG†

Summary—**In a new method of displaying the predicted paths and real-time positions of artificial earth satellites, cathode-ray tubes are used with two types of map overlays: orthographic projection and Mercator projection. Functions to match these projections are generated electronically. Auxiliary satellite information is cataloged and made available through "light gun" interrogation of the satellites displayed on the CRT. The display and orbit simulator can be used as an integral part of a live tracking system.**

A mathematical derivation of the subsatellite path on the orthographic projection is given.

INTRODUCTION

ELEVEN man-made objects are now orbiting in space around the earth. In the foreseeable future, an increasing number of satellites will undoubtedly be placed in orbit. As the frequency of launchings increases, so does the need for a panoramic visual display of all satellite positions in space relative to the earth's surface.

The repetitive nature of the orbits and the stability of the paths of the satellites through space point to a simple and reasonably accurate method of display. In the method under development at the Astrosurveillance Sciences Laboratory of the Air Force Cambridge Research Center, electronic analog simulators are used to generate the functions representing the predicted satellite orbits. The paths are presented visually on the face of a large-screen cathode-ray tube (CRT), over which is overlaid a transparent map projection of the earth. The predicted paths are generated at a frequency high enough to prevent flickering when presentation is time-shared. The real-time positions of the satellite are displayed as intensified spots on their respective traces and travel at rates corresponding to the speeds of the satellites.

After orbital characteristics have been established, the simulators must permit updating of the orbital paths by more accurate position information obtained from periodic observation of the satellite itself.

The system is sufficiently flexible to allow time to be moved forward and backward, up to one week. The value of this feature becomes apparent from the nature of the corrections that are possible in the simulated orbit. Although very accurate measurements of the real orbits are available from sophisticated sources, the data take time to process, and arrive late. If the system is turned back to the time when the observation was made, it is possible to inspect the simulated orbit, make any necessary correction, and then return the system to the present time. On occasions when it is desirable to know the satellite situation at some time in the future—as, for instance, if it is important to know when a communication or surveillance satellite will be over a particular area of the earth—time can be advanced until the simulated satellite position coincides with the specified position on the map overlay. The date and time can then be read off the calibrated controls that advance the time.

Access to other pertinent satellite data is accomplished through interrogation of the display with a "light gun." This is a photocell device that produces a video signal when pointed at the real-time spot of light on the CRT. Coincidence between the video signal and an electronic gate representing the sequenced time slot for displaying that specific satellite actuates a relay. The relay then controls a number of indicator circuits that present cataloged information relevant to the interrogated satellite, including identification, distance of apogee and perigee, date of launching, type of scientific and/or military data transmitted from the satellite, and maximum and minimum speeds. The same relay also samples a "range from earth" potentiometer geared to track with the real-time position generator, one indicator thus providing altitude readings for all satellites.

THE SATELLITE DISPLAY

Any presentation of satellite orbital information is inherently limited by the difficulty of accurately displaying a spherical area on a flat surface. Regardless of the type of map projection used, distortion is inevitable in the scale, perspective, and shape of the orbit. The rather difficult task of minimizing distortion to maintain a comprehensible, three-dimensional presentation then becomes a problem of the technical feasibility of compensating for the distortion.

Recent advances in display techniques may lead to substantially larger presentations. The newly developed mosaic displays that use gaseous, chemical, or electroluminescent cells have possibilities. A cathode-ray tube having a spherical screen—if it could be designed—would present the most realistic picture, but most requirements can be satisfied if the CRT has a hemispherical screen, which can be designed. To avoid delaying the display program, the present models are being designed around large-screen flat-faced CRT's because of their availability and standard deflection requirements.

A system of time-sharing shows all orbits in sequence above the CRT flicker rate. To display the predicted paths in time sequence, periodic functions that match the map projection are generated at a 5-kc frequency. This allows 100 targets to be displayed at a 50-cps rate.

The system's accuracy is governed primarily by the CRT and deflection system used in the display. Experi-

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ence has shown that in a conventional CRT and deflection system, errors can be held under 0.33 per cent. Analog techniques should therefore satisfy the accuracy requirements of the complete system. If the combined errors of the display and function generators are not greater than 0.5 per cent, positions on the earth traversed by the satellite will be displayed within 125 miles, corresponding to 36 seconds of orbit time for a 2-hour orbit. This accuracy should be adequate for most applications, since a satellite orbiting at an altitude of 1000 miles can theoretically be seen by a line-of-sight detector 3000 miles from the detection location.

Two different types of map projections are used in the displays currently under development in the Laboratory: orthographic and Mercator.

The Orthographic Projection

In the orthographic projection (perspective of the earth from infinity), the map overlay on the CRT is a hemisphere; two displays show the total earth. On this presentation, an orbital path projected onto a stationary earth (earth not rotating on its axis) appears as a portion of an ellipse. To show predicted paths relative to the earth's surface (Fig. 1), electronic function generators compensate for the distortion that the earth's rotation causes in the presentation of the elliptic pattern. For a polar projection, the sine and cosine functions of two oscillators of slightly different frequencies provide an electronic analog that satisfies the mathematical expressions.

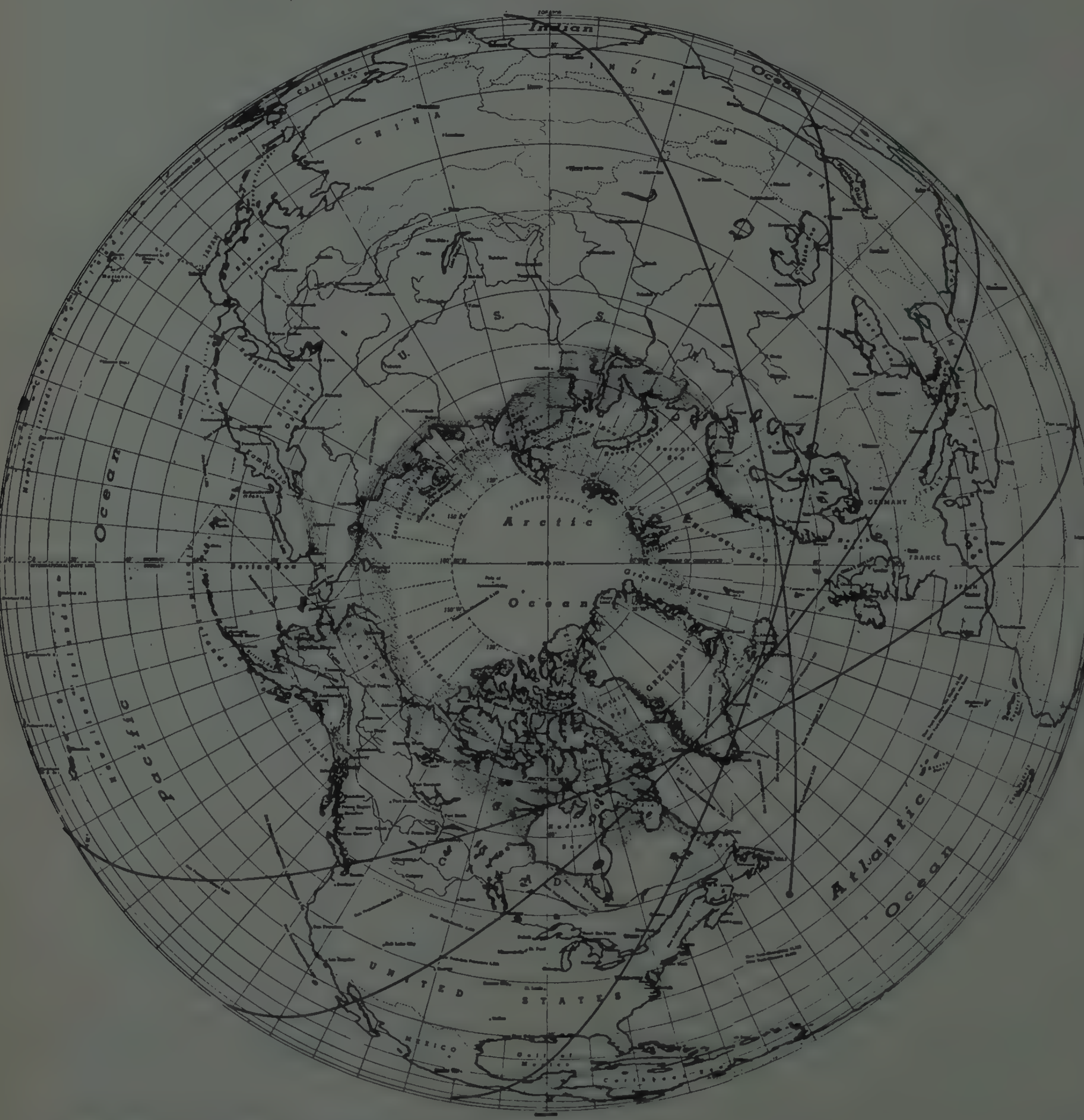

Fig. 1—Several orbits of an artificial earth satellite on an orthographic polar projection showing precessing due to earth's rotation.

Displayed on a CRT, the functions trace out the predicted path of the satellite. Each successive trace, in speeded-up time, precesses the same number of degrees that the earth will rotate under each successive satellite orbit.

The real-time motion and position of the satellite orbit is generated by a computer driven by a clock motor in real time. Sine and cosine potentiometers rotating at slightly different rates perform the same role as the electronic oscillators. The real-time position of the satellite is established on the display by repeatedly locking the start of the predicted trace to the x, y positions as determined by the real-time computer.

The locking gate can be manually controlled to show the predicted path of the satellite anywhere from zero time to several orbits. The CRT beam is held momentarily at the beginning of the sweep, thereby intensifying the real-time position of the satellite.

A MATHEMATICAL DERIVATION OF AN ORTHOGRAPHIC SUBSATELLITE TRACK

The following exercise shows the derivation of the mathematical equations of motion for the subsatellite track (projection of the orbit onto the earth's surface) to 1.0 per cent accuracy on an orthographic map projection.

The known equations of motion of a satellite in a plane are transformed to the equations of motion of a satellite in three dimensions with respect to axes fixed to the earth. As a final step, the equations of motion in three dimensions are converted to the subsatellite position as a function of time.

The mathematical exercise consists of the following steps:

1) The equations of motion for a set of axes whose x member coincides with the semimajor axis of the satellite's elliptic path are transformed to a new set of axes in the same orbital plane *but whose x member, instead of passing through the perigee, is so oriented as to pass through the point of maximum latitude of the orbit.* (The point of maximum latitude is shown in Fig. 4 as point M.)
2) The equations of motion are then referred to a set of axes that are fixed in space. The x and y members are in the plane of the earth's equator, with the z member perpendicular to the equatorial plane.
3) The equations of motion for a rotating system fixed to the earth are obtained. The result represents the position of the satellite in three-dimensional space, given in Cartesian coordinates as a function of time.
4) As a final step, the result is transformed to project the actual path in space onto the earth, thus forming the subsatellite equations of motion.

The following notation will apply in this section:

a = semimajor axis of ellipse
e = eccentricity of ellipse
i = orbit inclination to the equatorial plane
s = angular frequency of satellite $(2\pi/T_s)$
ω = argument of perigee
ω_e = angular frequency of earth's rotation $(2\pi/T_e)$
T_e = period of earth's rotation
T_p = time of perigee crossing
T_s = period of satellite

$$E = s(t-T_p) + e \sin s(t-T_p) + e^2/2! \sin 2s(t-T_p) + \cdots .$$

Equations of Motion

The path of a point mass moving in an inverse-square central force field is a conic section. Under the conditions assumed here, it will be an ellipse of low eccentricity (Fig. 2), one of whose foci is the center of the central force field (in this case the center of the earth).

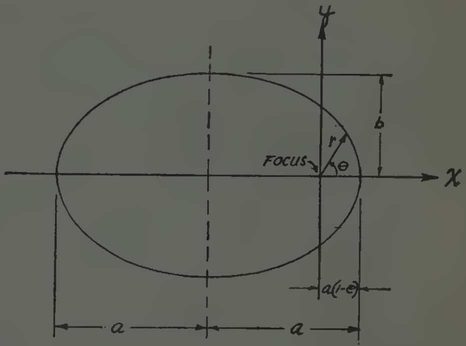

Fig. 2—Geometry of elliptic orbit. a = semimajor axis, b = semiminor axis, e = eccentricity

$$e = \frac{\sqrt{a^2 - b^2}}{a} .$$

According to Moulton, the path has the mathematical expression

$$r = \frac{a(1 - e^2)}{1 + e \cos \theta} . \tag{1}$$

Eq. (1) is an expression for the path, but what is required is a parametric representation, $y=y(t)$ and $x=x(t)$. The exact solution of this problem is, in general, impossible to obtain in closed form. The following equations are derived from Moulton:[1]

$$x = a(\cos E - e),$$
$$y = \sqrt{1 - e^2} \sin E. \tag{2}$$

If $e \leq 0.1$, only the first two terms of the infinite series for E are required to insure better than 1 per cent accuracy in $\sin E$ and $\cos E$ terms.

---

[1] F. R. Moulton, "An Introduction to Celestial Mechanics," The Macmillan Co., New York, N. Y., pp. 164–165; 1947.

Transformation I

Transformation I is illustrated in Fig. 3. The orientation of the coordinate system with respect to the earth is given in Fig. 4.

Elementary coordinate transformations give the following relationships between the x, y coordinate system and the x', y' coordinate system. (Note that $z \equiv Z' \equiv 0$, the satellite being assumed to move in the x, y or x', y' plane.)

$$\begin{aligned} x' &= x \cos \beta + y \sin \beta; \\ y' &= -x \sin \beta + y \cos \beta. \end{aligned} \tag{3}$$

Since $\omega = 90° - \beta$, (3) can be rewritten as

$$\begin{aligned} x' &= x \sin \omega + y \cos \omega; \\ y' &= -x \cos \omega + y \sin \omega. \end{aligned} \tag{4}$$

For this analysis, ω is assumed constant, even though perturbation effects cause the ellipse to precess slightly.

Transformation II

Transformation II yields the equations of motion with respect to a set of axes fixed in space, with x and y members in the equatorial plane. Note that the mathematical rotation of axes is around the y' axis, thus making $y' \equiv y''$ (Fig. 5).

Elementary coordinate transformation equations give

$$\begin{aligned} x'' &= x' \cos i - z' \sin i, \\ y'' &\equiv y', \\ z'' &= x' \sin i + z' \cos i. \end{aligned} \tag{5}$$

Transformation III

The system of coordinates x'', y'', z'' is now referred to a rotating earth. The coordinate system designated by x''', y''', z''' represents the satellite position in space. Transformation III (Fig. 6) gives

$$\begin{aligned} x''' &= x'' \cos (\gamma + \omega_e t) + y'' \sin (\gamma + \omega_e t), \\ y''' &= -x'' \sin (\gamma + \omega_e t) + y'' \cos (\gamma + \omega_e)t, \\ z''' &\equiv z''. \end{aligned} \tag{6}$$

Solving (2) through (6) simultaneously for x''', y''', and z''' gives

$$\begin{aligned} x''' = {} & A \cos (E + \omega_e t + \theta) \\ & + B \cos (E - \omega_e t + \phi) + C \cos (\omega_e t + \psi), \\ y''' = {} & D \cos (E + \omega_e t + \delta) \\ & + F \cos (E - \omega_e t + \alpha) + G \cos (\omega_e t + \rho), \\ z''' = {} & H \cos (E + \tau) - e k_5; \end{aligned} \tag{7}$$

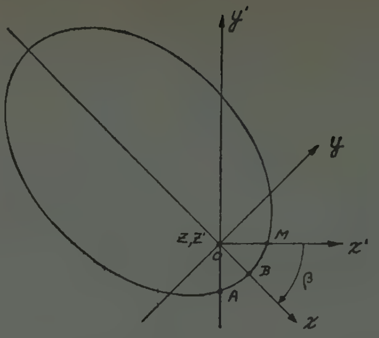

Fig. 3—Transformation I. Z and Z' are perpendicular to the page, directed positively outward.

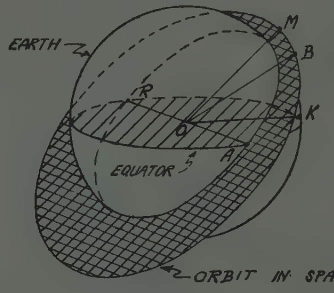

Fig. 4—Orientation of coordinate system in Fig. 3. M = point of maximum latitude, B = perigee, K = point on equator, the same longitude as point M, β = angle MOB, ω = argument of perigee angle BOA, angle MOA is 90°, angle β is $90° - \omega$, i = inclination of orbit to equator, angle MOK.

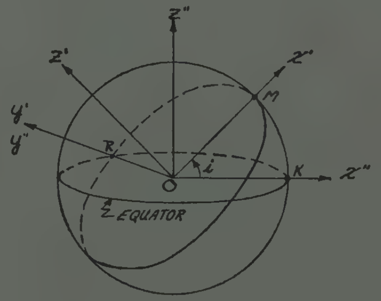

Fig. 5—Rotation of axes by Transformation II. $Z' \perp x'$ and y'. i = inclination of orbit.

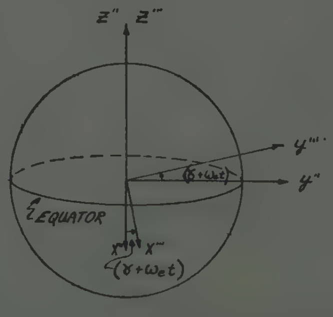

Fig. 6—Transformation III. x'' and y'' are in the plane of the equator but fixed in space. x''' and y''' are in the plane of the equator but fixed to earth.

where

$$
\begin{aligned}
A&=\tfrac{1}{2}\sqrt{(k_1-k_4)^2+(k_2-k_3)^2}, & \theta&=\gamma-\tan^{-1}\frac{(k_2-k_3)}{(k_1-k_4)};\\
B&=\tfrac{1}{2}\sqrt{(k_1+k_4)^2+(k_3+k_2)^2}, & \phi&=\gamma-\tan^{-1}\frac{(k_3+k_2)}{(k_1+k_4)};\\
C&=\tfrac{e}{2}\sqrt{(k_3)^2+(k_1)^2}, & \psi&=\gamma-\tan^{-1}\frac{(k_3)}{(-k_1)};\\
D&=\tfrac{1}{2}\sqrt{(k_2-k_3)^2+(k_4-k_1)^2} & \delta&=\gamma-\tan^{-1}\frac{(k_4-k_1)}{(k_2-k_3)};\\
F&=\tfrac{1}{2}\sqrt{(k_2+k_3)^2+(k_4+k_1)^2}, & \alpha&=\gamma-\tan^{-1}\frac{(k_4+k_1)}{-(k_2+k_3)};\\
G&=e\sqrt{(k_3)^2+(k_1)^2}, & \rho&=\gamma-\tan^{-1}\frac{(k_3)}{(k_2)};\\
H&=\sqrt{(k_5)^2+(k_6)^2}, & \tau&=-\tan^{-1}\frac{(k_6)}{(k_5)};
\end{aligned}
\tag{8}
$$

where

$$
\begin{aligned}
k_1 &= a\sin\omega\cos i; & k_2 &= a\sqrt{1-e^2}\cos\omega\cos i;\\
k_3 &= a\cos\omega; & k_4 &= a\sqrt{1-e^2}\sin\omega;\\
k_5 &= a\sin\omega\sin i; & k_6 &= a\sqrt{1-e^2}\cos\omega\sin i.
\end{aligned}
\tag{9}
$$

Transformation IV

With the equations of motion in three-dimensional space determined, it is now necessary to transform these relations to the equations of motion of the subsatellite track, in a form amenable to electronic simulation. Transformation IV is illustrated in Fig. 7. The coordinates of the subsatellite track are

$$x_{ss}=\frac{R}{r}x'''; \qquad y_{ss}=\frac{R}{r}y'''; \qquad z_{ss}=\frac{R}{r}z'''. \tag{10}$$

Solving for x_{ss}, using Moulton's equation $r=a(1-e\cos E)$, gives

$$x_{ss}=\frac{Rx'''}{a(1-e\cos E)}. \tag{11}$$

If $e\leq 0.1$, (11) may be approximated with only the first two terms of the Taylor series, as

$$x_{ss}\approx\frac{R}{a}(1+e\cos E)x'''. \tag{12}$$

Substituting the expression for x''' in (7) into (12) gives

$$
\begin{aligned}
x_{ss}=\frac{R}{a}(1+e\cos E)\big[&A\cos(E+\omega_e t+\theta)\\
&+B\cos(E-\omega_e t+\phi)+C\cos(\omega_e t+\psi)\big].
\end{aligned}
\tag{13}
$$

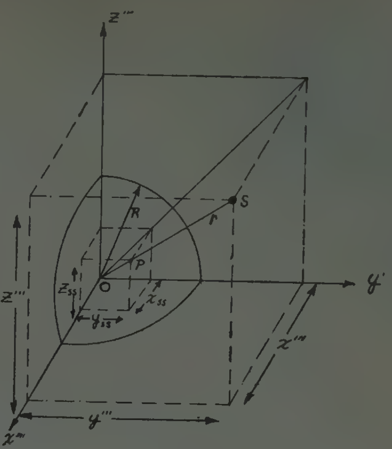

Fig. 7—Transformation IV. r is the distance of the satellite from the center of the earth, R is the radius of the earth and P is the projection of the satellite (s) onto the earth.

Multiplied out,

$$
\begin{aligned}
x_{ss}=\frac{R}{a}\big[&A\cos(E+\omega_e t+\theta)+B\cos(E-\omega_e t+\phi)\\
&+C\cos(\omega_e t+\psi)\big]+\frac{Re}{a}\big[A\cos E\cos(E+\omega_e t+\theta)\\
&+B\cos E\cos(E-\omega_e t+\phi)\\
&+C\cos E\cos(\omega_e t+\psi)\big].
\end{aligned}
\tag{14}
$$

The approximation

$$E=s(t-T_p)+e\sin s(t-T_p) \tag{15}$$

for $e\leq 0.1$ per cent, produces errors not greater than 1 per cent in sin E and cos E terms.

Assuming that errors no greater than 1 per cent can be tolerated (15), various trigonometric relations, and approximations based on a Taylor series, can be used to reduce (14) to

$$
\begin{aligned}
x_{ss}=\frac{R}{a}\big\{&A\cos[(s+\omega_e)t+\theta-sT_p]\\
&+B\cos[(s-\omega_e)t+\phi-sT_p]\\
&+eA\cos[(2s+\omega_e)t+\theta-2sT_p]\\
&+eB\cos[(2s-\omega_e)t+\phi-2sT_p]\\
&+C\cos(\omega_e t+\psi)\big\}.
\end{aligned}
\tag{16}
$$

To calculate y_{ss}, it is not necessary to carry out the full procedure used in solving for x_{ss}, but merely to determine how x''' and y''' are related. Inspection shows that when

$$
\begin{aligned}
&x_{ss}\to y_{ss}, && x''\to y''';\\
&A\to D, \quad B\to F, && C\to G;\\
&\theta\to\delta, \quad \phi\to\alpha, && \psi\to\rho.
\end{aligned}
$$

Therefore,

$$\begin{aligned} y_{ss} = \frac{R}{a} & D\cos\left[(s+\omega_e)t + (\delta - sT_p)\right] \\ & + F\cos\left[(s-\omega_e)t + (\alpha - sT_p)\right] \\ & + De\cos\left[(2s+\omega_e)t + (\delta - 2sT_p)\right] \\ & + Fe\cos\left[(2s-\omega_e)t + (\alpha - 2sT_p)\right] \\ & + G\cos(\omega_e t + \rho)\big\}. \qquad (17) \end{aligned}$$

To solve for z_{ss}, the procedure is the same as for x_{ss}. Thus,

$$z_{ss} = \frac{Rz'''}{a(1 - e\cos E)} \approx \frac{R}{a}(1 + e\cos E)z'''. \qquad (18)$$

Substituting from (7), (18) gives

$$z_{ss} = \frac{R}{a}(1 + e\cos E)\left[H\cos(E + \tau) - ek_5\right]. \qquad (19)$$

Simplifying, as in the case of x_{ss}, gives

$$\begin{aligned} z_{ss} = \frac{R}{a} & \left\{H\cos\left[s(t - T_p) + \tau\right]\right. \\ & \left. + eH\cos\left[2s(t - T_p) + \tau\right] - ek_5\right\}. \qquad (20) \end{aligned}$$

Circular Orbit, a Special Case

For a circular orbit $e=0$, the equations become greatly simplified. Where r is the radius of the orbit,

$$\begin{aligned} x_{ss} &= \frac{R}{r}\left\{A\cos\left[(s+\omega_e)t + \theta - sT_p\right]\right. \\ &\qquad \left. + B\cos\left[(s-\omega_e)t + \phi - sT_p\right]\right\}; \\ y_{ss} &= \frac{R}{r}\left\{D\cos\left[(s+\omega_e)t + \delta - sT_p\right]\right. \\ &\qquad \left. + F\cos\left[(s-\omega_e)t + \alpha - sT_p\right]\right\}; \\ z_{ss} &= \frac{R}{r}\left\{H\cos\left[s(t - T_p) + \tau\right]\right\}. \qquad (21) \end{aligned}$$

To consider a specific case, assume that $\omega=0$, $T_p=0$, and $\gamma=0$. Then

$$k_1 = k_4 = k_5 = 0,$$

$$k_2 = r\cos i, \qquad k_3 = r, \qquad k_6 = r\sin i;$$

and

$$A = \frac{r}{2}(1 - \cos i) = r\sin^2\left(\frac{i}{2}\right),$$

$$B = \frac{r}{2}(1 + \cos i) = r\cos^2\left(\frac{i}{2}\right),$$

$$C = 0,\ D = A,\ F = B,\ G = 0,\ H = r\sin i;$$

$$\theta = -90°,\ \phi = -90°,\ \psi = +90°,\ \delta = 0,$$

$$\alpha = 0,\ \rho = -90°,\ \tau = -90°.$$

Substituting these values in (21) gives

$$\begin{aligned} x_{ss} &= R\left\{\sin^2\left(\frac{i}{2}\right)\cos\left[(s+\omega_e)t - 90°\right]\right. \\ &\qquad \left. + \cos^2\left(\frac{i}{2}\right)\cos\left[(s-\omega_e)t - 90°\right]\right\}, \\ y_{ss} &= R\left\{\sin^2\left(\frac{i}{2}\right)\cos\left[(s+\omega_e)t\right]\right. \\ &\qquad \left. + \cos^2\left(\frac{i}{2}\right)\cos(s-\omega_e)t\right\}, \\ z_{ss} &= (R\sin i)\cos(st - 90°); \qquad (22) \end{aligned}$$

or,

$$\begin{aligned} x_{ss} &= R\left\{\sin^2\left(\frac{i}{2}\right)\sin\left[(s+\omega_e)t\right]\right. \\ &\qquad \left. + \cos^2\left(\frac{i}{2}\right)\sin(s-\omega_e)t\right\}, \\ y_{ss} &= R\left\{\sin^2\left(\frac{i}{2}\right)\cos\left[(s+\omega_e)t\right]\right. \\ &\qquad \left. + \cos^2\left(\frac{i}{2}\right)\cos(s-\omega_e)t\right\}, \\ z_{ss} &= (R\sin i)\sin st. \end{aligned}$$

It is evident that three frequencies must be generated to simulate a circular orbit; for a polar projection (one of the poles in the center of the display), only two frequencies are required.

For the special case where $i=0$, the equations reveal what is intuitively apparent: $z_{ss}=0$, meaning that the simulated satellite is moving along the equator and travels with an angular velocity equal to the difference between the satellite's angular velocity and the earth's angular velocity.

Another very interesting case is that of the 24-hour communications satellite, for which $\omega_e=s$. For $i=0$, it is apparent that $x_{ss}=\text{constant}$, $y_{ss}=\text{constant}$, and $z_{ss}=0$. Because of the particular choice of ω and γ, $x_{ss}=0$, $y_{ss}=R$. An additional modification of this case is when $i\neq 0$. Inspection of the z_{ss} and x_{ss} axes reveals that $z_{ss}(t)$ and $x_{ss}(t)$ are parametric equations for a figure-eight Lissajous pattern.

The foregoing exercise has demonstrated the derivation of the equations of motion of the subsatellite track as explicit functions of time and known constants of the orbit. It must be remembered, however, that the results are only approximate to within 1 per cent for satellites with eccentricity not greater than 0.1. For satellites of higher eccentricity, the mathematical analysis is much more complex and does not appear to be amenable to a practical direct electrical analog simulation. It may be possible to cancel some of this complexity through special electronic function generators designed for curve-fitting.

THE MERCATOR PROJECTION

In the second display method, a Mercator map projection presents the earth's geographic coordinates as an x, y grid. This projection can display the total area of the earth and has, therefore, been used extensively for plotting satellite courses (Fig. 11).

A Mercator projection spreads the area at the 80th parallel so that the east-west direction near the poles has the same dimension as the east-west direction at the equator. The system consequently produces a continually varying mileage scale, a land mass the size of Greenland seeming to have approximately the same area as South America. Because of this distortion, the electronic display voltages produced by the simulated-orbit generators are not amenable to a reasonable mathematical derivation. A great circle (circular orbit) *on* the equator appears as a straight line traveling east and west. Great circles *close* to the equator appear as sinusoids which have progressively larger amplitude and distortion the farther north and south they go. As the orbit approaches either pole, the sinusoid flattens off. A great circle that goes through the poles must travel parallel to the longitudinal lines and so appears as a square wave.

The orbits of most of the earth-circling satellites launched to date can be simulated for display on the Mercator projection by sine waves modified in a diode function generator, but the orbit of Explorer VI, launched in August, 1959, requires a different approach. The path of this orbit, although elliptic in space, takes on a peculiar shape when presented on a Mercator projection (Fig. 8). This irregularity is due to the relation between the satellite's speed and the earth's rotational speed. At perigee, approximately 260 km from earth, the satellite's speed is approximately 23,000 mph. As the satellite coasts out to apogee, 42,000 km away, it slows down to 2200 mph. Since any point on the earth at the equator is traveling at a constant 1000 mph, Explorer VI's east-west speed component projected onto the earth is sometimes faster and sometimes slower than the earth's rotational speed.

In a plot of the satellite course $y=f(x)$ on a Mercator map (Cartesian coordinate system), there are segments where y is a multivalued function of x. Since the x coordinate represents display time on the CRT, and the spot cannot be in more than one position at one time, the original deflection circuit had to be revised. The curve can be represented by two parametric equations $y=f_1(t)$; $x=f_2(t)$. Then x and y are single-valued functions of time that can be simulated in electronic function generators and applied to the x, y deflection circuit of the CRT (Fig. 9). The beam of the CRT traces out the desired function, and time is measured along the trace.

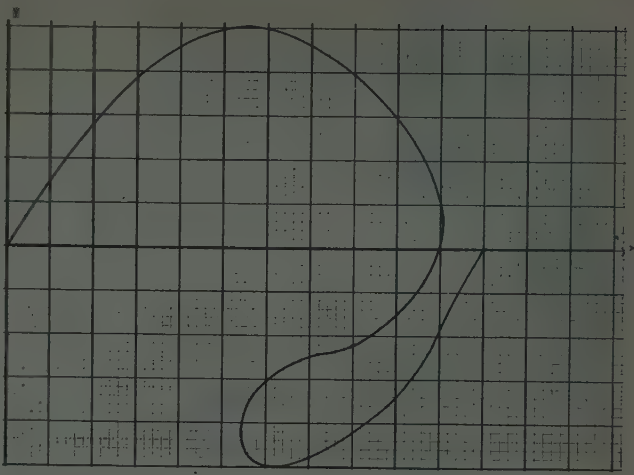

Fig. 8—Path of Explorer VI satellite on a Mercator projection. Note that y is a multivalued function of x (display time).

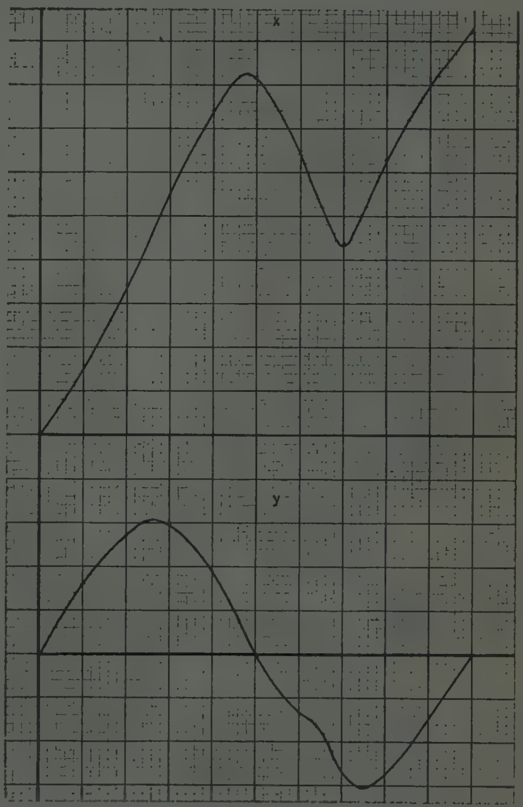

Fig. 9—Explorer VI orbit represented by two parametric curves $y=f_1(t)$; $x=f_2(t)$.

Bright spots moving along the predicted paths indicate the positions of the satellites in real time. The motion of the satellites is generated electromechanically, with a synchronous motor driving adjustable speed reducers coupled to potentiometers to produce the proper orbits as slowly varying dc voltages. The dc voltages are compared with linear sawteeth produced in the 5-kc time slots allotted to the respective orbits (see Fig. 10).

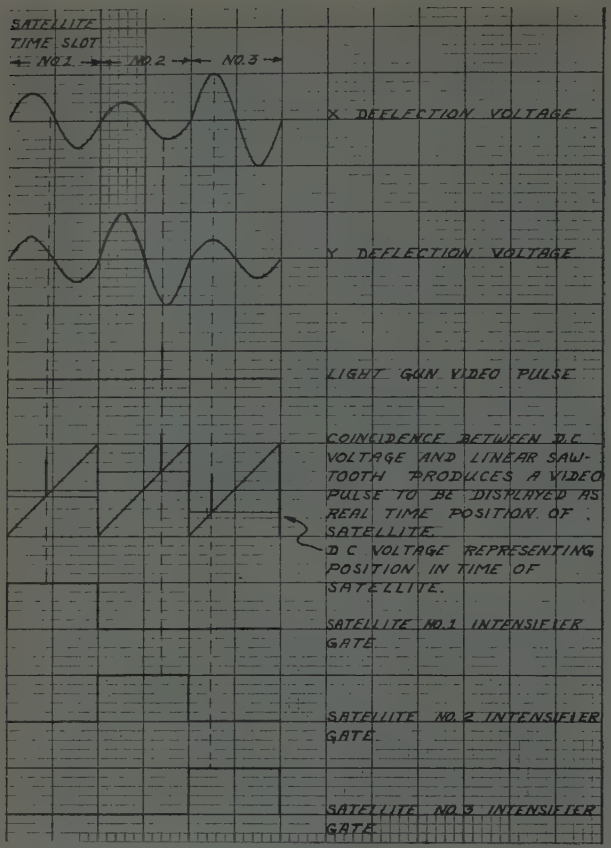

Fig. 10—Cathode-ray tube timing waveforms.

Coincidence between a sawtooth voltage and a dc voltage generates a video timing pulse that is displayed as a bright spot (Fig. 11) on the trace of the orbit and indicates the instantaneous position of the satellite.

Since the speed of any satellite in a noncircular orbit is continually varying, maximum at perigee and minimum at apogee, accommodation for the changing speeds is provided through a suitable cam in the mechanical portion of the real-time position generator.

Because precessing of the orbit projected onto the earth is a function of the satellite's orbital frequency in relation to the earth's rotational frequency, it must be individually compensated for in each case. When the real-time position has completed 360° on the Mercator projection, the predicted trace is shifted the appropriate number of degrees for the start of the next cycle. This is accomplished by automatic shifting of a timing gate that selects the phase of the x, y functions to be displayed and clamps them to the proper starting voltage.

Display of Height Information

Height, the third component, is programmed into the simulator and made available on command. For each satellite, a linear height potentiometer mechanically tracks the real-time position. A removable cam mechanically linked to the potentiometer provides a function that converts the linear voltage to the actual height voltage. A digital voltmeter can be automatically connected to read the instantaneous height of any satellite interrogated by the light gun. When the height profile has changed sufficiently to warrant it, a new cam is inserted to make the readings more accurate.

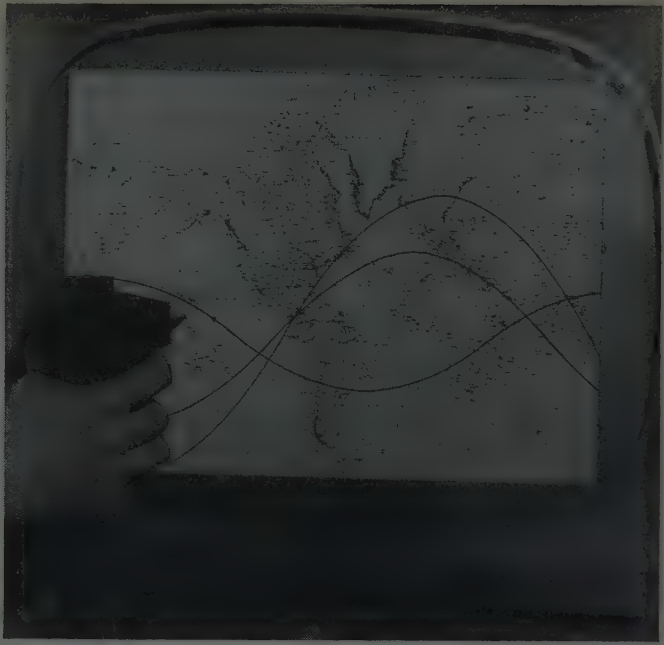

Fig. 11—"Light gun" interrogating satellite displayed as a bright spot on CRT trace with Mercator overlay.

The deviation of the earth's shape from a true sphere produces two additional effects that influence tracking and display: one, a slight precession of the orbit in the direction of the earth's rotation; the other, a similarly small precession of the orbit in its own plane. In other words, the apogee and perigee are continually rotating. These are minor perturbations of a few degrees per day, but they must be compensated for in the timing mechanism.

Because the timing of the complete system is controlled by a clock, it is possible to turn time backward and forward and inspect the satellite situation in the past and future as well as in the present.

Display of Auxiliary Information

Auxiliary satellite data can be stored in several ways. One would be to use recorded voice, with a separate channel for each satellite. Another would be to use a film that would project the information onto a screen. The simplest way, adequate for the present, is the use of identification lights that, when activated, refer to a file card on which the information is typed.

Access to the information is by means of a "light gun" (Fig. 11). This technique utilizes correlation between the time-sequenced voltage waveforms generated for displaying the positions of satellites and the pulse generated by the light gun when it is pointed at the satellite's position on the CRT. Coincidence between the two signals activates a relay that initiates the display action dictated by the particular storage scheme used. The following satellite information is cataloged:

1) identification,
2) date and time of launching,
3) size and weight,
4) height of apogee and perigee,
5) maximum and minimum speed,
6) inclination of orbit to equator,
7) type of scientific and/or military data transmitted from satellite,
8) pertinent ancillary information.

In addition to interrogating targets on the display, it is possible to locate the track and position of any specific satellite by operating an identification switch on the panel. This action intensifies the trace of that particular orbit, thereby distinguishing it from the others. Through control switches, it is also possible to select the orbits to be displayed. In the future, when the number of satellites has increased to the point where displaying them all at once might not be feasible, it may be desirable to select satellites for display according to category, such as communications, weather, or surveillance.

SIMULATOR USED AS AN AUTOMATIC TRACKING SYSTEM

Future developments will include the capability of making automatic corrections in the simulated orbits from live data at the detection sites. Receiving periodic corrections on all satellites, the unit could be considered an important component in a live satellite detection and tracking system.

Since the electronic function generators scan all positions of the satellites as represented on the map overlay, it is possible to sample the functions with the real-time position video signals and store the instantaneous position voltages on capacitors. The stored dc voltages would represent the latitudes and longitudes of each satellite. With appropriate gating circuits allowing the generated functions of each satellite orbit to be corrected only by the live data representing their respective satellites, the device remotely resembles a track-while-scan unit used in radar systems. Application of this technique to the automatic tracking of satellites is possible because of the repetitive nature and stability of the orbits in space.

ACKNOWLEDGMENT

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The Navy Space Surveillance System*

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***Summary*—A complete system for satellite detection and tracking and for computations of satellite orbits has been built by the Navy under ARPA sponsorship. This detection system uses a CW transmitter separated from two receiving sites, all having fan-type coplanar antenna beams. The angle of arrival of the reflected signals is measured at each station by the use of an interferometer. The position of the reflecting object is inferred by the point in the fan antenna beam defined by the intersection of the arrival angles at the two receiving stations.**

Two ARPA-sponsored Space Surveillance radal (radio detection and location) devices of the type described have been installed in the southern U. S. In addition to the detecting and tracking installation the system includes data transmission lines, a data reduction center, a very high speed computer for orbit determination and predictions, and display devices.

INTRODUCTION

THE PROBLEM of detecting satellites is made difficult by their great height and velocity, and small size. The high velocity implies a short observation time but also implies great trajectory stability, so that a small number of independent observations properly separated on the ellipse serves to determine the orbit. The problem of detecting satellites is much different from that of tracking aircraft and ships for which the pulse radar was designed. For this earlier problem the target can maneuver rapidly and therefore required continuous tracking.

The detection system is shown schematically in Fig. 1. Here a transmitter antenna provides illumination in a narrow fan beam which is coplanar with similar re-

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† U. S. Naval Research Lab., Washington 25, D. C.

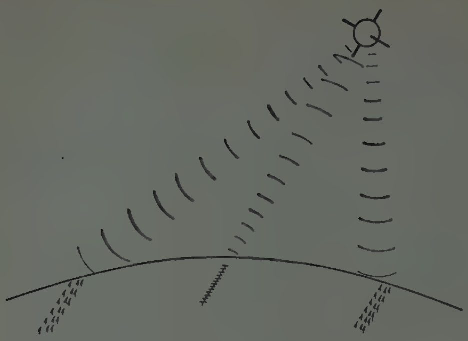

Fig. 1—The detection system used for space surveillance.

ceiving station beams. (Only that portion of the fan beam striking the satellite is shown in the figure.) The position of a reflecting object in the common antenna patterns is inferred by measuring the angle of reception of the reflected signals by means of interferometers at the two receiving sites.

The operation of the receiving stations is similar to that used by the system known as Minitrack[1] initiated by M. Rosen for use with Project Vanguard. The technique he suggested was similar to that used by several missile tracking systems, by radio astronomers, and by sonar. Minitrack was based largely on the missile tracking system AZUSA, built by the Convair Division of the General Dynamics Corporation. The operational principles of both systems appear to have been first conceived by Dr. H. L. Saxton of this Laboratory.[2]

Another contribution was made by Dr. J. J. Freeman[3] who analyzed this particular type of system. From his analysis it was simple to show that an interferometer system operating at a frequency of about 100 mc would require a radiated power of only a few milliwatts.

One of the problems existing in such a system is that of calibrating the angle measured electrically to the local zenith angle. Two techniques were implemented to provide the required calibration. One, involving ballistic camera photography of an airplane-carried flashing light, was demonstrated as practical by E. Habib, then of this Laboratory, and has been used as the standard calibration method for this system. The other method involved reflecting a signal off the moon. Then by knowing the local zenith angles of the moon and the corresponding electrical readings the required calibration was obtained, though with an accuracy much reduced from that obtainable with the ballistic camera.

The moon reflecting system consisted of a 50-kw FM transmitter installed to operate into a 50-ft dish at the Signal Corps Engineering Laboratories (now USARDL) at Ft. Monmouth, N. J. The transmitter was located and the work of the two laboratories was coordinated through the good offices of Lt. Col. E. J. Hagerman, Signal Corp Liaison Officer at the Naval Research Laboratory.

The moon reflection system was placed in operation in December, 1957. Early in 1958 the transmitter at SCEL was used to provide reflected signals from the remaining Sputnik (1957 Beta) into the tracking station at Blossom Point, Md., as shown in Fig. 2.

On June 20, 1958, the Advanced Research Projects Agency of the Department of Defense authorized the Naval Research Laboratory to provide one tracking complex in the Eastern United States and one in the Western United States as shown in Fig. 3. The NRL system is being expanded to form a continuous line across the Southern United States.

The first portion of the system to be placed in operation involved the stations at Fort Stewart, Ga., and Jordan Lake, Ala. This portion of the system was put in operation on July 29, 1958—less than six weeks after the ARPA Order was issued. This schedule was met by transporting the transmitter from SCEL to Jordan Lake and by modifying the Fort Stewart Minitrack station antennas to be compatible with this new system.

The Silver Lake installation became operational in November, 1958 and the Western Complex in February, 1959. The electronic equipment was built by Bendix Radio, the antennas by the Technical Appliance Company, and the transmitters were furnished by Multronics, Inc. The stations are operated by Bendix Field Operations Personnel.

The Receiving System

The receiving system has been described previously[1] so it will be discussed only briefly. Fig. 4 shows the principle of operation. After the signals from the two antennas are first amplified by preamplifiers, the output can be divided as necessary for the various phase comparisons required. By means of local oscillator circuits, the signals from the two antennas can be amplified in a single channel, thereby reducing the differential phase shift. When the combined signal is detected and compared to the difference frequency, the resulting phase is equal to the phase difference between the signals arriving at the two antennas. The zero reading of the phase between the two signals will not normally read zero for an equal phase signal at the antenna. The difference between the phase read and the phase present at the antennas is found by means of a system of calibration.

Calibration

As was mentioned, an interferometer system requires calibration and the most accurate means of calibration is the ballistic camera method. With the size of antennas used for space surveillance, however, the far zone of the antenna is many miles high—too high for conventional aircraft.

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[1] J. T. Mengel, "Tracking the earth satellite, and data transmission, by radio," Proc. IRE, vol. 44, pp. 755–760; June, 1956.

[2] H. L. Saxton, NRL Rept. No. 4003, July, 1952. (Confidential.)

[3] J. J. Freeman, "Principles of Noise," John Wiley and Sons, Inc., New York, N. Y., pp. 271–274; 1958.

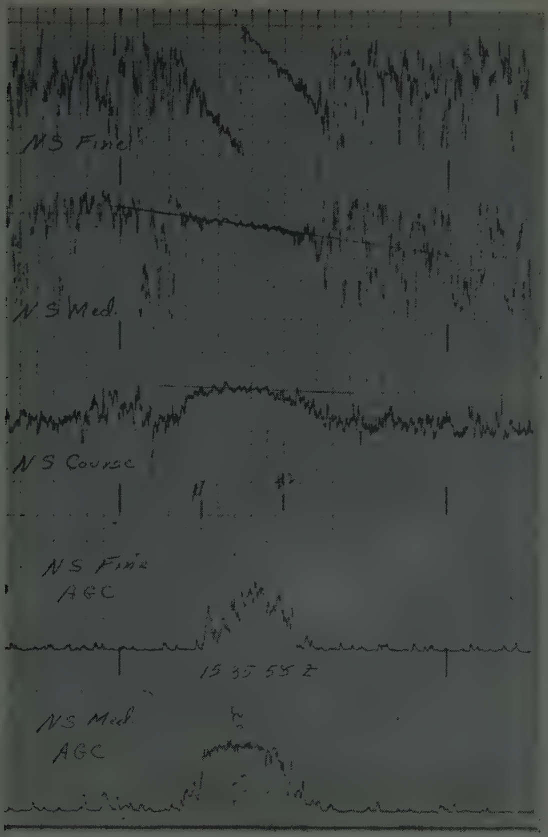

Fig. 2—Reflected signal obtained from 1957 Beta satellite.

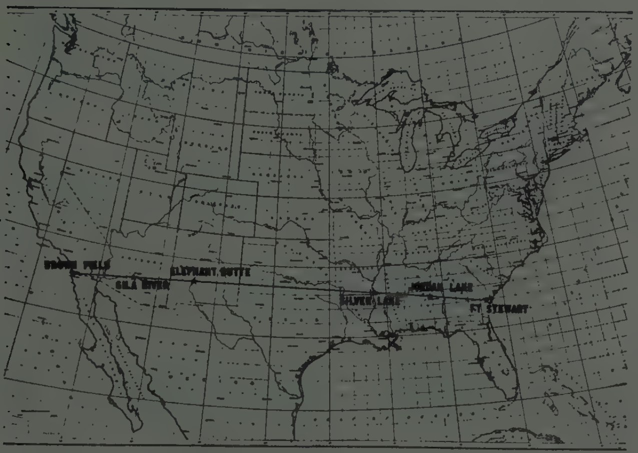

Fig. 3—Location of tracking stations.

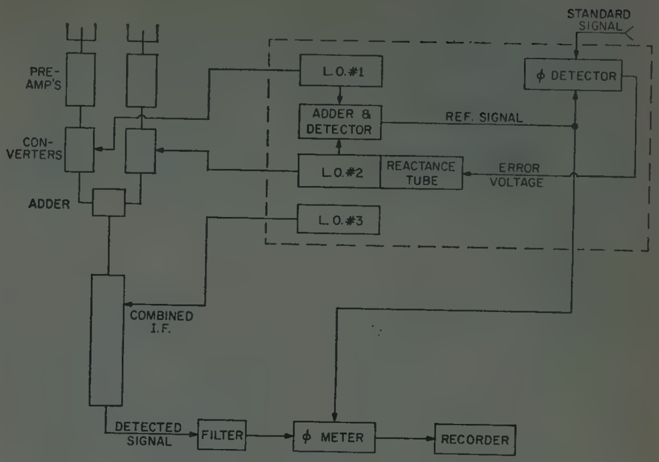

Fig. 4—Block diagram of receiving system.

For calibration a distant source in a known position is required. A device that meets this requirement is the Vanguard I satellite, 1958 Beta, whose position is well known from the orbit computed by NASA. Even though the orbit of this satellite is the best known of any, the data for calibration purposes must still be obtained from post flight rather than predicted data. Once the data is obtained the calibration is made by comparison with the electrical data. Since the orbital data is obtained by calibrating the Minitrack stations against the stars, the Space Surveillance stations are also calibrated against the same standards—though with diminished accuracy.

SIGNALS

Fig. 5 shows the signal obtained from the calibration satellite, 1958 Beta. The inclination of this satellite is nearly the same as the latitude of the Space Surveillance line, so its signals remain in the antenna beamwidth for a considerable time. In general the signal reflected will appear similar to a very small portion of this direct signal, though the reflected signals are much noisier than the direct signals. A reflected signal from another satellite is shown in Fig. 6.

In addition to signals reflected from satellites many extraneous signals are present. They appear as reflections from meteorite trails, from aircraft, and as direct signals from radiating satellites, electrical storms, radio stars, direct feedthrough from transmitter to receiver, and man-made interference. One of the most prevalent interfering signals is due to meteorites. Fig. 7 shows one of the signals obtained during a time of high meteorite activity.

The simplest means of reducing meteorite reflections is illustrated in Fig. 8. Here the polarization of the transmitting antenna is opposite to that used on the receiving antenna. The record also shows a receiving antenna having the same polarization as the transmitter. A count of incidences on the two receiving channels shows that approximately only 5 per cent as many signals appear on the receiving antennas having the opposite polarization to the transmitter as appear on the antenna having the same polarization as the transmitter.

The reason for the reduction in meteorite reflections is understood when it is realized that the meteorite trails in general are formed below the ionosphere and that the most usual excitation of the meteorite trail is the one due to the indivudual electrons oscillating in the polarization of the exciting element. Since the excitation of the transmitting antenna is not subject to Faraday

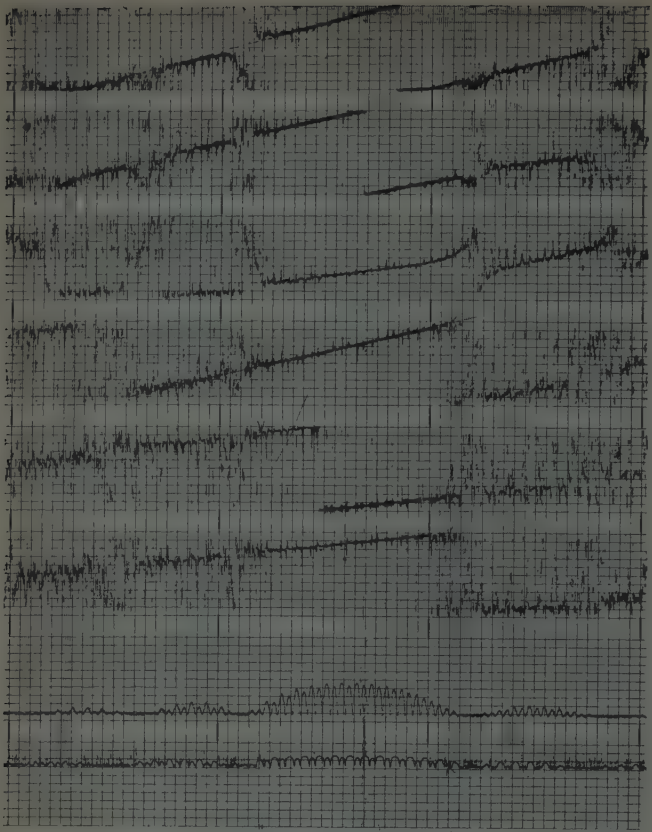

Fig. 5—Direct signal from 1958 Beta 2 (Vanguard I) satellite.

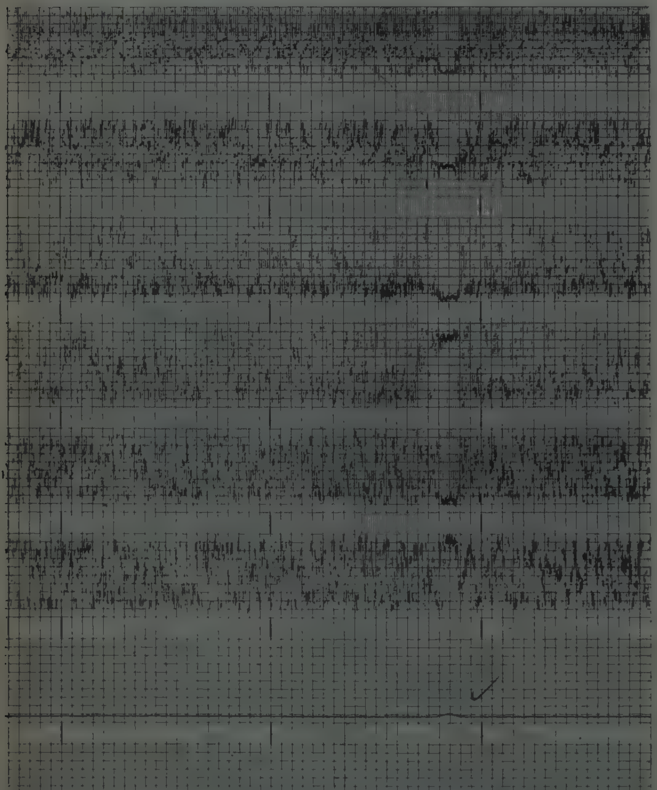

Fig. 6—Reflected signal from satellite.

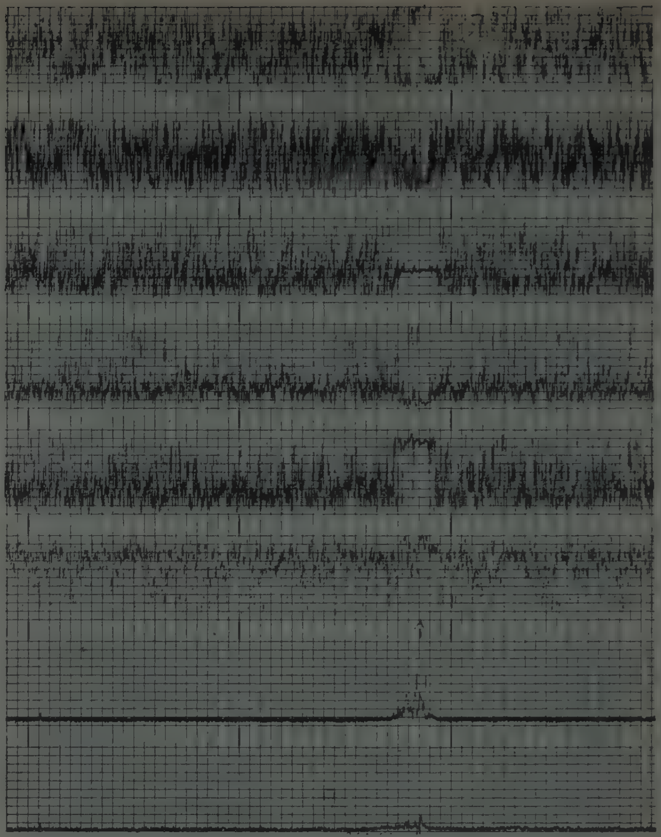

Fig. 7—Reflected signal from meteorite trail.

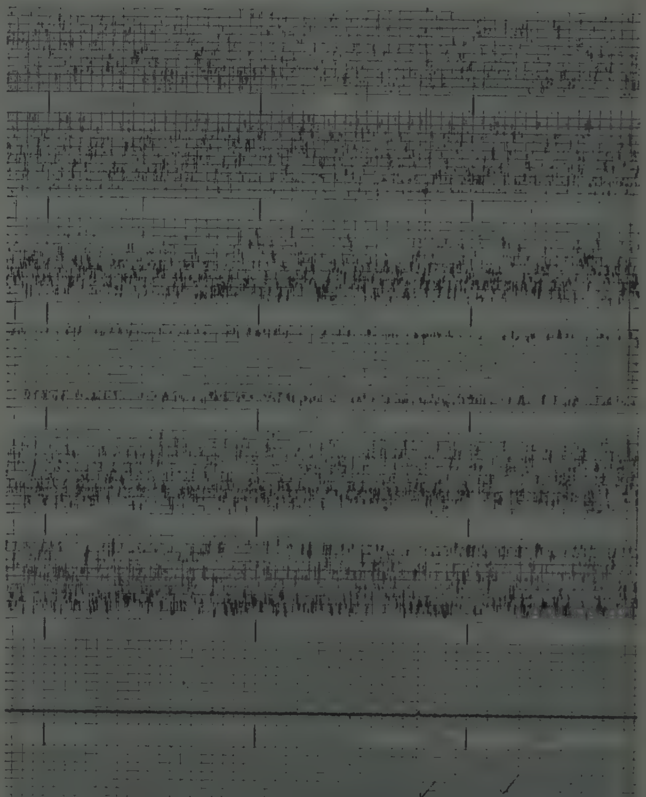

Fig. 8—Reduction in reflections due to meteorite trails by cross polarizing transmitting and receiving antennas. (Bottom trace shows the output of a receiver having the same antenna polarization as the transmitter; the next trace above shows the output of a receiver having the antenna polarization crossed to that of transmitter.)

rotation at the usual height of meteorite trails, the type of reflection described above will be cross-polarized to an antenna having a polarization crossed to the transmitter. Under these conditions the response from meteorites can be expected to be low. Reflections from satellites occur from within or above the ionosphere and the polarization of such signals is subject to Faraday rotation so the polarization of the received signal can be expected to be random.

STATIONS

The stations involve large antenna installations. The Jordan Lake transmitting site is shown in Fig. 9. The Fort Stewart installation is considered the R&D receiving station and hence is larger than the other receiving sites. At each receiving station the information from the phase meters is placed on direct reading recorders and is also placed on phone lines for real time transmission to the Naval Research Laboratory and to the Naval Weapons Laboratory at Dahlgren, Va. The data is sent using standard FM telemetry subcarriers directly on a standard telephone line.

SPACE SURVEILLANCE OPERATIONS CENTER

During the experimental phase of system operation, the Space Surveillance Control Center is located at NRL. It is normally operated on an 8-hour day and 5-day week. An experimental operations center is also located there both for the development of improved operational techniques and for serving as a backup for the Space Surveillance Operations Center at Dahlgren, Va., which is in operation around-the-clock seven days a week. The analog data representing phase and signal-level information of interest are read manually immediately after they are recorded at Dahlgren. The observations are used for improving the predictions of known satellites and determining the orbits of unknown satellites crossing the Space Surveillance line.

The Space Surveillance Operations Center has been established at the Naval Weapons Laboratory (Dahlgren, Va.) so that it could take advantage of the NORC (Naval Ordnance Research Computer) and of the NWL staff. The NORC computer is especially well suited for orbit computations since it has both a high precision (13 significant decimal digits plus sign and exponent) and a high speed (15,000 operations per second). A picture of the NORC is shown in Fig. 10 and a table of its operating characteristics is as follows:

Number system:	Decimal (Automatic floating point).
Word size:	16 digits and check digit.
Instructions:	3-address.
Storage capacity:	2000 words (electrostatic). 20,000 words (magnetic core) being added.
Storage access time:	8 μsec.
Multiplication time:	31 μsec.
Addition time:	15 μsec.
Magnetic tape units:	8 operating at 70,000 characters per second.
Printers:	2 mechanical operating at 300 characters per second. 1 optical printer operating at 15,000 characters per second.

Fig. 9—The Jordan Lake transmitting site.

Fig. 10—The Naval Ordnance Research Computer (NORC).

ORBIT COMPUTATIONS

The mathematical formulation of the orbit-computation programs has been the responsibility of Dr. P. Herget (Director of the Cincinnati Observatory) under contract to NRL and of Dr. G. M. Clemence and Dr. R. L. Duncombe, both of the U. S. Naval Observatory. The analysis, programming, and some of the mathematical formulation are being done by NWL.

The orbit computations required for satellites are divided into two parts: a) determination of the orbital elements from the observations, and b) prediction of future satellite positions from these elements (compilation of an ephemeris). For initial orbit determination and prediction, Cowell's method of numerical integration[4] is used; for large numbers of observations, the method of general oblateness perturbations[5] is employed. Six elements are needed to describe an elliptical orbit at some given time T_0 (called the epoch). These elements are derived from the observations when there is sufficient information available to be equivalent to a position vector and a velocity vector at a given time.

The orbital elements are refined automatically on the NORC by successive application of a differential correction procedure until the residuals between all the observations and the positions computed from the elements have become small. The best available set of elements is used to compute a position vector in geocentric inertial space for the satellite for every minute of time in the future (for days or weeks, depending on the type or orbit) for predictions. Both in orbit determination and prediction, the position of the satellite as computed from the elements is corrected for perturbations due to the earth's oblateness and atmospheric drag. Depending on the requirements of the user, predictions as needed are computed in the form of a world map (longitude and latitude of the subsatellite point on the earth and the height of the satellite above the earth), or of local station predictions (range, bearing, and elevation information at frequent time intervals for optical instruments or narrow-beam radars to view the satellite), or of beam-crossing predictions (time, height above the earth, and zenith angle) for Space Surveillance stations.

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[4] P. Herget, "The Computation of Orbits," 1958. (Published by the author.)

[5] P. Herget and P. Munsen, "A modified Hansen lunar theory for artificial satellites," *Astronomical J.*, vol. 63, pp. 430–433; November, 1958.

The six elements needed to describe an elliptical orbit at time T_0 (epoch) may be given as follows:

Semimajor axis (a)
Eccentricity (e)
Inclination (i)
Right ascension of ascending node (θ_0)
Argument of perigee (ω_0)
Mean anomaly at epoch (M_0).

The epoch is given as the year, month, day, hour, minute, and second in universal time. The semimajor axis of the ellipse is measured in earth's equatorial radii (equatorial radius is 3963.34 statute miles based on the international ellipsoid for the shape of the earth[6]). The eccentricity is a ratio which is always less than unity for an ellipse and is equal to zero for a circle. The inclination (see Fig. 11) is the angle measured in degrees from the equatorial plane of the earth to the orbital plane at the ascending node (that is, where the satellite crosses the equator in a northward direction). The right ascension of the ascending node is measured in degrees eastward along the earth's equator from the vernal equinox to the ascending node. The vernal equinox establishes the x-axis for the geocentric inertial coordinate system, the z-axis passing through the earth's North pole, and the y-axis lying in the equatorial plane to form a right-handed coordinate system. The argument of perigee (point of closest approach to the earth) is the angle in degrees measured in the orbital plane from the ascending node to the perigee. The mean anomaly at epoch is given in degrees and represents the position of the satellite in the orbit with respect to the perigee.

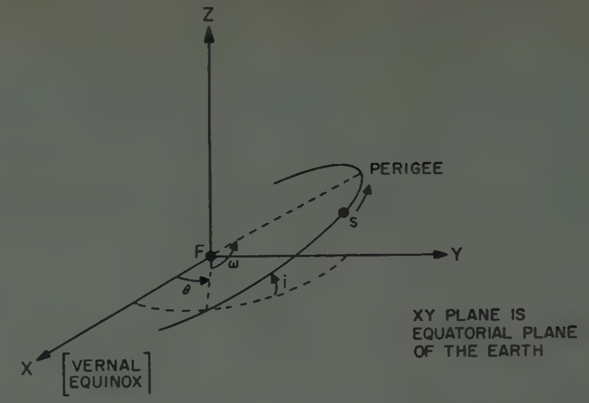

Fig. 11—Coordinate reference system.

If there were no disturbing forces exerted on the satellite, its orbit would not change with time and predictions of future satellite positions would be made simply on the basis of an ellipse fixed in inertial space. For most artificial earth satellites, there are two major disturbing forces which must be taken into account: one resulting from the shape of the earth (the equatorial bulge), and one resulting from the presence of the atmosphere. Because of the distribution of the mass of the earth, there is motion of the perigee in the orbital plane and motion of the node in the equatorial plane. Because of the energy removed from the orbit by atmospheric drag on the satellite, the orbit collapses with time, the apogee (farthest point from the earth) shrinking many times faster than the perigee. These perturbations are taken into account both in orbit determination and prediction.

An example of a complete set of elements and related information is given for the satellite 1958 Beta 2 (Vanguard I) as released from the Vanguard Computing Center as follows:

Epoch	October 22, 1959	122700 UT
Anomalistic period	134.04899	Minutes
Period decay	−0.0001	Minutes per day
Inclination	34.249	Degrees
R.A. of ascending node	191.613	Degrees
Motion of node	−3.023	Degrees per day
Argument of perigee	182.671	Degrees
Motion of perigee	4.415	Degrees per day
Latitude of perigee	−1.503	Degrees
Mean anomaly at epoch	160.459	Degrees
Eccentricity	0.18963	
Semimajor axis	1.36030	Earth radii
Perigee height	405.6	Statute miles
Apogee height	2,450.3	Statute miles
Velocity at perigee	18,371	Miles per hour
Velocity at apogee	12,514	Miles per hour

The anomalistic period is the time for the satellite to make a complete revolution from perigee to perigee. Depending on the orbital elements, the motions of node and perigee are only a few degrees a day at most. For a polar orbit ($i=90°$), there is no motion of the node; and for an orbit having an inclination of $63\frac{1}{2}°$, there is no motion of the perigee. The perigee and apogee heights are measured from the surface of the earth, which is taken to correspond to the equatorial radius in this case.

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[6] "The American Ephemeris and Nautical Almanac for the Year 1959," U. S. Naval Observatory, Washington, D. C.; 1957.

A catalog of all satellites is maintained by the NORC computer and can be displayed as subsatellite positions projected on a world map. Fig. 12 illustrates the appearance of the display, called SPASCORE, for which the film is normally projected at one frame per minute in "real time" but may be run forward or backward at 20 frames per second. The map overlay on the projection screen is a modified cylindrical projection of the earth. The subsatellite positions are calculated by the NORC and automatically recorded on 35-mm film from a Charactron cathode-ray-tube output device, which is also used as an optical printer for alphanumeric data output. This equipment provides great versatility in displaying the computer output; it can be operated on-line with the computer to project the output in real-time with a delay of as little as eight seconds or to record the output on film at 40 frames per second for later development and projection.

The most important output of the Space Surveillance System is information on nonradiating satellites. Orbital elements can be computed and predictions of future satellite positions can be given to those authorized to receive such information. The System also has been used to augment the observations of radiating satellites.

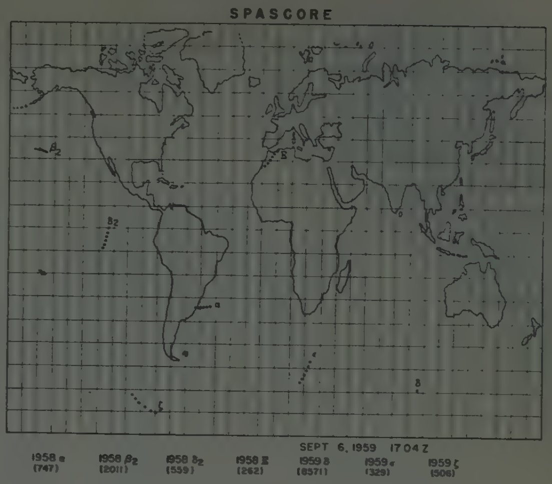

Fig. 12—NORC output display (SPASCORE).

Discussion

The radal (radio detection and location) system described differs from a pulse radar in principle and in detail differs greatly. Pulse radar[7] uses a single installation from which energy is transmitted and received. The location of reflecting objects is inferred from the angle of arrival and from the time delay between transmission and reception. Previous to and during World War II the angle of reception was derived from the antenna direction. The monopulse radar,[8] also developed at this Laboratory, derives its angular information from several simultaneous observations of the angle of reception of the incident wave. Pulse radars were developed for detecting and tracking ships and aircraft, objects which have great maneuverability but low speeds.

The radal system is designed to detect and locate objects having great speed but very limited maneuverability. For such objects a great range capability but a modest number of sightings serves to determine the path of the object for days to come.

To satisfy the requirement of detection at great ranges the system has been designed to maximize range capability. Since a larger average power can be generated economically at CW than with pulses, a CW system is indicated. To use antennas having large capture areas without unusably small beamwidths, a low frequency is used. The antenna beamwidths are designed to detect over a large angle in one direction and a very narrow angle in the other. This technique permits detection of objects passing through an area of large dimensions but small volume.

An integral part of the Space Surveillance System is a high-precision, high-speed digital computer to process the observations with a minimum time delay. Thus a catalog of all orbiting objects within range of the system is maintained to produce predictions of their future positions for the use of authorized customers.

The present system has demonstrated itself to be effective and reliable. Improvements are planned and will be installed as time and funds allow. An additional transmitter is expected to be located in the central portion of the line in the near future. The present system of analog data transmission and manual data reading is expected to be replaced by a digital transmission system to permit fully automatic data handling by the computer.

Acknowledgment

The authors wish to acknowledge their indebtedness to Dr. C. E. Cleeton, Superintendent of the Applications Research Division, and Captain W. E. Berg, Senior Program Officer for Military Applications of Satellites, U. S. Naval Research Laboratory, Washington, D. C., for their continuing helpful guidance on the Space Surveillance project, which has engaged the talents and interest of a large part of the Division. Above all, they wish to acknowledge the tireless efforts of their colleagues, to whom they attribute the success of the project.

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[7] A. H. Taylor, "Radio Reminiscences: a Half Century," Naval Research Lab., Washington, D. C., pp. 294–308; 1948.

[8] R. M. Page, "Monopulse radar," 1955 IRE Convention Record, pt. 8; pp. 132–134.

Ionospheric Scintillations of Satellite Signals*

H. P. HUTCHINSON†, SENIOR MEMBER, IRE AND P. R. ARENDT†

Summary—**The scintillation of satellite-emitted radio signals has been observed using two different techniques, namely, Doppler-shift frequency measurements and radio direction finding. In this shortened version of our complete paper, there are given the results obtained using Doppler-shift measurements only. Variations from a smooth Doppler-shift curve obtained during individual orbits give a measure of the frequency scintillation occurring and thus of the roughness of the ionospheric path between the satellite and the observer. As expected, these variations are a function of frequency, and they become less as the frequency is increased.**

THE Doppler shift that is observed is primarily a function of the radial velocity of the satellite with respect to the observer and the emitted frequency. However, for frequencies in the region between 20 and a few hundred mc, the effects of the wave velocity of propagation through the ionosphere as a function of the refractive index are noticeable. Therefore, the observed Doppler shift is further affected by changes in refractive index or ionization along the path between the satellite's position and that of the observer. Also the Doppler shift is affected by the intensity of ionization at the satellite. The degree to which the observed variations are attributable either to changes in ionization at the satellite when the satellite is passing through the ionosphere or to changes in the integral of electron density along the path, is a suitable subject for further study.

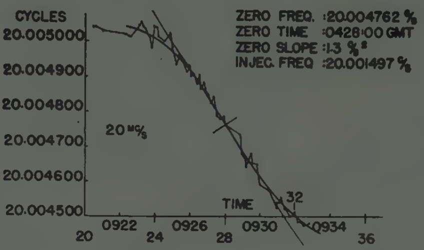

Fig. 1—Satellite δ 1958, orbit #2294, October 29, 1958.

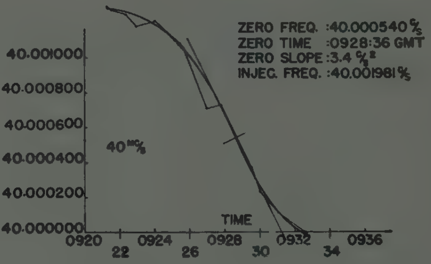

Fig. 2—Satellite δ 1958, orbit #2294, October 29, 1958.

These frequency measurements were taken with an instrumental accuracy of ±7 cycles in frequency measurement and at 1-second time intervals based on simultaneous reception of WWV standard time transmissions. Fig. 1 shows the deviations of the observed Doppler-shift recordings of Sputnik III from a smooth curve drawn through the observed points. It is seen that substantial variations occur even during the point of nearest approach of the satellite. Fig. 2 shows the variations observed with the first harmonic on the same orbit, namely for 40 mc. It is seen that even at this frequency substantial scintillations occur as deviations from a smooth curve. Now, going to 108 mc we see a fairly smooth curve described by observations of Explorer I shown in Fig. 3. The center part is quite smooth but, as expected from a longer ionospheric path at the outer ends of the curve, these ends show some deviations from a smooth curve. This orbit has an apogee of 2500 km and a perigee of 360 km. Thus, for a good part of the time it lies within the ionosphere, and the ionospheric path from the observer to the satellite is somewhat less than the total thickness of the ionosphere in the direction of the satellite.

Such is not the case for the orbit of Vanguard I, which at all times is well above the maximum of the normal *F* region. Its orbit has an apogee of 3960 km and a perigee of 660 km. Fig. 4 shows frequency deviations of the order of 20 to 25 cycles observed during a particular orbit of Vanguard I. This orbit occurred during a time when the ionospheric quality figure was 7 (good). Subsequently, on the next day, there occurred the largest short-period deviation, which we have observed on the 108-mc signals (Fig. 5). This anomoly also was taken from an orbit of Vanguard I. It appears to be of greater magnitude than what might be expected from variations in the integrated electron density of the ionosphere. The quality index was "good" again. Since the orbit of this satellite is well above the normal *F* region, and at the time of this observation the satellite was at an altitude of approximately 3900 km, that is, close to its apogee, a possibility exists that the occurrence may have been caused by the satellite encountering a cloud of charged particles above the ionosphere. However, one must also consider the possibility of an ionized cloud coming between the satellite's position and the observer for the three-minute time duration of the anomoly by some disturbance in the *F* region of the ionosphere.

From this and the previous illustrations, it is clear that ionospheric frequency scintillations of substantial magnitudes do occur with satellite signals. This observation coincides with the so-called "color" scintillations

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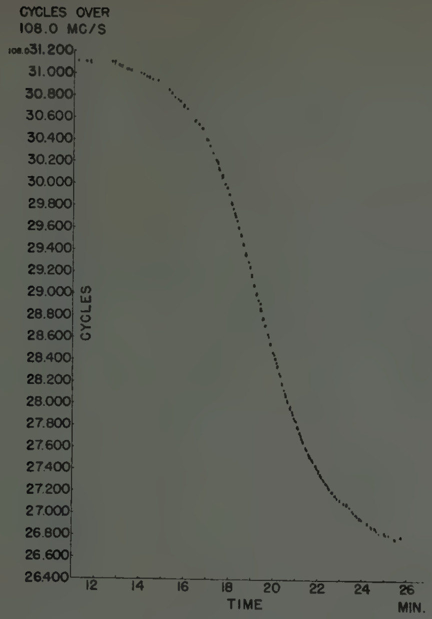

Fig. 3—Satellite 1958 α, orbit #97, February 8, 1958. Time reference 1811:00 EST.

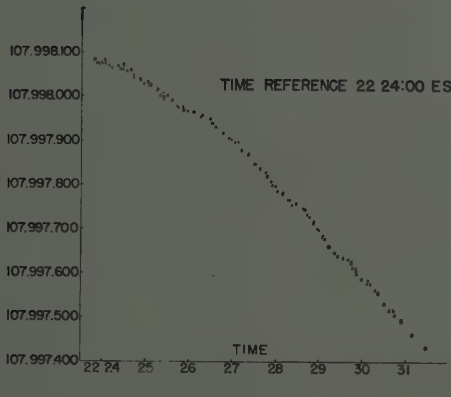

Fig. 4—Satellite β 1958, orbit #128, March 28, 1958.

of the radio star signals which have been cited by Little.[1] With the recent public announcement of the Argus high-altitude nuclear detonations and the Christofilos effect, our Doppler-shift recordings were studied to see whether any systematic effect was discernible. As a preliminary measurement, the records of eighty-nine orbits were examined visually and ranked on a qualitative grading basis from 0 to 5 with the number zero denoting no scintillations and the number five, maximum scintillations. We call this a scintillation quality index. These observations were all on Sputnik III orbits and taken at 20 mc, which is our most sensitive frequency for that purpose. Fig. 6 shows histograms of these distributions taken before, during and after the three Argus high-altitude detonations. It is clearly seen that systematic changes occurred which initially showed a strong reduction in scintillation of the Doppler-shift signals, and even several days later indicated a much lower index than for the period before the detonations. Meanwhile, quantitative measurements were initiated using the measured areas generated by the difference between a smoothed curve through the data and the curve generated by the data points. For each orbit this area was measured in arbitrary units and the resultant number was divided by the time (in minutes) of the observation. The results of this procedure are shown in Fig. 7, where data to the left of the first vertical line cover the period of 18 days prior to the first detonation, and to the right, the period of 16 days afterwards. It is clear that in this latter period the scintillations of these Sputnik III signals were systematically less than prior to the event.

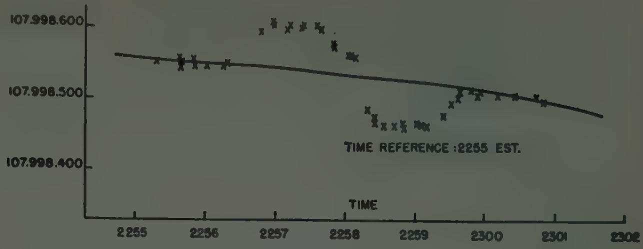

Fig. 5—Satellite β 1958, orbit #138, March 29, 1958.

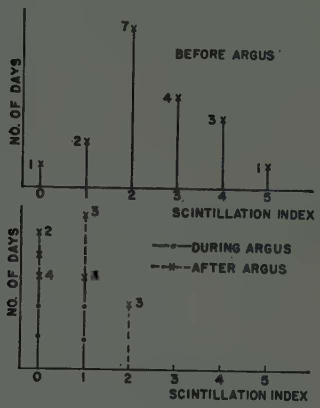

Fig. 6—Distribution of scintillation quality index.

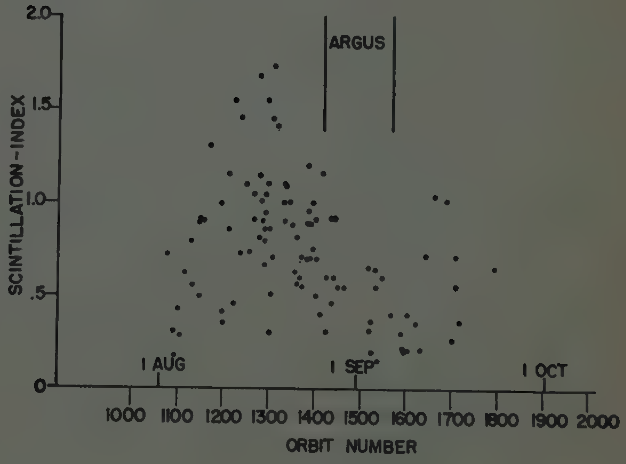

Fig. 7—Delta '58 Sputnik III.

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[1] G. Little, "The Jodrell Bank program on meteors, aurora, and ionosphere," *J. Geophys. Res.*, vol. 59, pp. 152–155; March, 1954.

The Satellite Ionization Phenomenon*

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Summary—A number of observations are presented which show a close correlation between CW-reflected HF signals and passes of artificial earth satellites. The periodic (nonrandom) occurrence of the signal bursts and the symmetry of some burst sequences are indicative of satellite-related phenomena. The occurrence of a variety of satellite-related Doppler effects are described and several satellite ionization mechanisms are also discussed. The possible relation of the satellite phenomenon to prior solar activity is mentioned.

INTRODUCTION

IN an earlier communication,[1] it was suggested that some anomalies in the signals from Sputnik I, noted at The Ohio State University Radio Observatory in October, 1957, might be caused by satellite-induced ionization. Subsequent experiments using CW reflection provided evidence that artificial earth satellites could, on occasion, be detected due to such ionization.[2,3] The CW-reflection technique was the same as employed earlier by Wylie and Castillo[4] for observing meteors, a standard short-wave receiver being used to monitor the transmissions of WWV, near Washington, D. C., on 20 mc.

During the day, ionospheric ionization is usually sufficient to reflect a strong signal from Washington to Columbus on this frequency. However, at night and in particular between midnight and 6 A.M. local time, the signals from WWV become extremely weak or inaudible except for brief signal bursts which appear to be associated with temporary localized increases in ionization in the ionosphere. Although many of the signal bursts were probably due to meteors, the marked periodicity of many of the longer and stronger bursts suggested that they were satellite related, many occurring close to the time of near approach of Sputniks I or II. Bursts caused by meteors would be expected to occur in time in a random manner; hence, it appeared highly probable that mechanisms other than meteor ionization were involved.

Many CW-reflection observations have been made subsequently which show effects related to other satellites. As an example, signal bursts observed on some of the first passes of Sputnik III are presented in the next section. The following section describes CW-reflection signal bursts observed during the next few weeks which can be interpreted as due to the systematic separation of the Sputnik III objects because of their differing drag characteristics. Later sections describe burst periodicities of very short period (order of 1 minute), symmetrical burst sequences, Doppler and other phenomena which appear to be satellite related, and a number of mechanisms which may account for the satellite-induced ionization.

OBSERVATIONS OF SPUTNIK III ON ITS FIRST PASSES

An interesting series of observations was made soon after Sputnik III was launched on May 15, 1958. At 7:19 A.M. EST, a WWV signal burst was recorded as Sputnik III crossed western Europe on the completion of its second trip around the earth. This burst is shown in the top record in Fig. 1, which is a photograph of the original Esterline-Angus tape made at the time. Time increases from right to left.

One hour and 47 minutes later, or at 9:06 A.M., a group of weak WWV bursts was recorded (second record from top in Fig. 1). At this time, Sputnik III was completing its third trip and was over the middle of the Atlantic Ocean. Another hour and 47 minutes later, or at 10:53 A.M., a series of strong WWV bursts was recorded as Sputnik III was completing its fourth trip and passing northeastward off the east coast of the United States (third record from top in Fig. 1). One hour and 47 minutes later, a fourth series of strong WWV bursts plus background enhancement were recorded as Sputnik III completed its fifth trip, over the central part of the United States (fourth record from top in Fig. 1). On the following day (May 16, 1958), during the Sputnik III pass over the eastern United States at 11:35 A.M., a series of strong WWV bursts was recorded (bottom record in Fig. 1).

The bars under the lower three records in Fig. 1 indicate the approximate times of near approach of the satellite as determined by observations of the Sputnik III transmissions on 20.005 mc. All deflections of the pen showing on the records for these passes were monitored by an operator and identified as being due to WWV (20.000 mc). Parts of the records are missing because at these times the receiver beat frequency oscillator (BFO) was turned on.

The WWV burst groups corresponding closely to the passes of Sputnik III at 10:53 A.M. and 12:40 P.M. on

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[1] J. D. Kraus and J. S. Albus, "A note on some signal characteristics of Sputnik I," PROC. IRE, vol. 46, pp. 610–611; March, 1958.

[2] J. D. Kraus, "Detection of Sputnik I and II by CW reflection," PROC. IRE, vol. 46, pp. 611–612; March, 1958.

[3] J. D. Kraus, R. C. Higgy, and J. S. Albus, "Observations of the U. S. Satellites Explorers I and III by CW reflection," PROC. IRE, vol. 46, p. 1534; August, 1958.

[4] L. R. Wylie and H. T. Castillo, "Clustering of meteors as detected by the use of radio technique," *Ohio J. Science*, vol. 56, pp. 339–347; November, 1956.

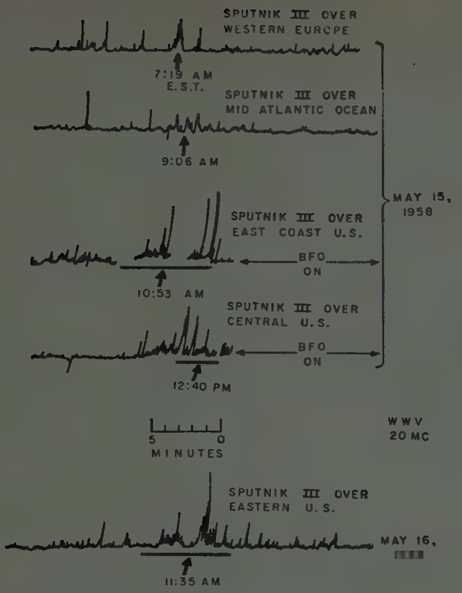

Fig. 1—Records of WWV 20-mc signals obtained during the first passes of Sputnik III on May 15 and 16, 1958. The times of near approach of the satellite are indicated by the heavy arrows.

May 15 and at 11:35 A.M. on May 16 are a striking phenomenon. It is highly improbable that these burst groups were caused by anything other than Sputnik III or effects associated with it. At that time, the instrumented satellite of Sputnik III, its large rocket carrier and several associated pieces of hardware were all close together in orbit. These components had only been in orbit for a short while and might conceivably have been degassing heavily, since they were operating in a nearly perfect vacuum and only shortly before had been at full atmospheric pressure.

The bursts at 7:19 and 9:06 A.M. on May 15 are not as strong, and although they occurred at the time of nearest approach of the satellite, their relation to the satellite is less certain.

Ordinarily, WWV puts a very strong signal into Columbus on 20 mc after 6 or 8 A.M. (EST). However, on the mornings of both May 15 and 16, the WWV level was very weak and close to the level ordinarily observed earlier in the morning between 12 and 6 A.M.

OBSERVATIONS OF SPUTNIK III OBJECTS BETWEEN MAY 26 AND JUNE 7, 1958

During the period a few weeks after Sputnik III was put into orbit, many records of the WWV 20-mc signals recorded at The Ohio State University Radio Observatory showed a large burst occurring consistently before the near approach of the Sputnik III instrumented satellite (1958 Delta 2). Of particular interest was the fact that the burst did not occur a constant number of minutes before the near approach of the satellite, but rather the time increased by a few minutes each day. A number of records of WWV 20-mc signals between May 26 and June 7, 1958 illustrating this effect are presented in Fig. 2. The characteristic 4-minute off periods of WWV from the 45th to the 49th minute of each hour may be noted on a number of the records. All records are so arranged that the time of nearest approach of the instrumented satellite is above the 0 minute point (scale at bottom). These times were determined by observations of the satellite's 20.005-mc transmissions by means of other receivers.

The receiver selectivity was sufficiently sharp that the satellite 20.005-mc transmissions do not show on the records in Fig. 2. At the right, a large WWV burst may be noted which occurs progressively earlier (further to the right) each day with respect to the time of nearest approach of the instrumented satellite. On May 26, the burst was about 30 minutes before the instrumented satellite and by June 7, it was nearly 70 minutes before. A line through these bursts extrapolated back intersects the zero line (for the instrumented satellite) at about May 15, which was the launching date, thus suggesting that this object was coincident with the instrumented satellite on that date. The line giving the approximate passage times for the rocket carrier (1958 Delta 1) determined from other data is indicated at the upper right of Fig. 2. Hence, if the moving series of bursts was due to an orbiting object, it could not have been either the instrumented satellite or the rocket carrier. Perhaps this object was one of the other pieces of hardware (nose cone or side shields) separated from these satellites at the time when they were put into orbit. Its period was obviously longer than that of the rocket but less than that of the instrumented satellite.

It is to be noted in Fig. 2 that large WWV bursts occurred on nearly all of the records at or somewhat before the time of nearest approach of the instrumented satellite (0 line). However, these bursts are not as large as for the earlier object. A further point of interest is that relatively few bursts were detected which appeared to be associated with the large rocket carrier. Hence, if these observations are significant, it may be concluded that the size of an artificial satellite is not necessarily the most important factor in producing detectable ionization.

The increasing separation between the bursts in Fig. 2 appears to be due to the systematic separation of the Sputnik III objects because of their differing drag characteristics. These observations are reminiscent of the burst groups detected during the terminal phase of Sputnik I due presumably to the gradual separation of a number of fragments of the satellite.[5]

BURST GROUPS OF 1-MINUTE PERIOD

As mentioned earlier, a marked periodicity in the WWV signal bursts may be taken as evidence of ionization induced by mechanisms other than meteors. Thus, burst periodicities corresponding to the orbital period

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[5] J. D. Kraus and E. E. Dreese, "Sputnik I's last days in orbit," PROC. IRE, vol. 46, pp. 1580–1587; September, 1958.

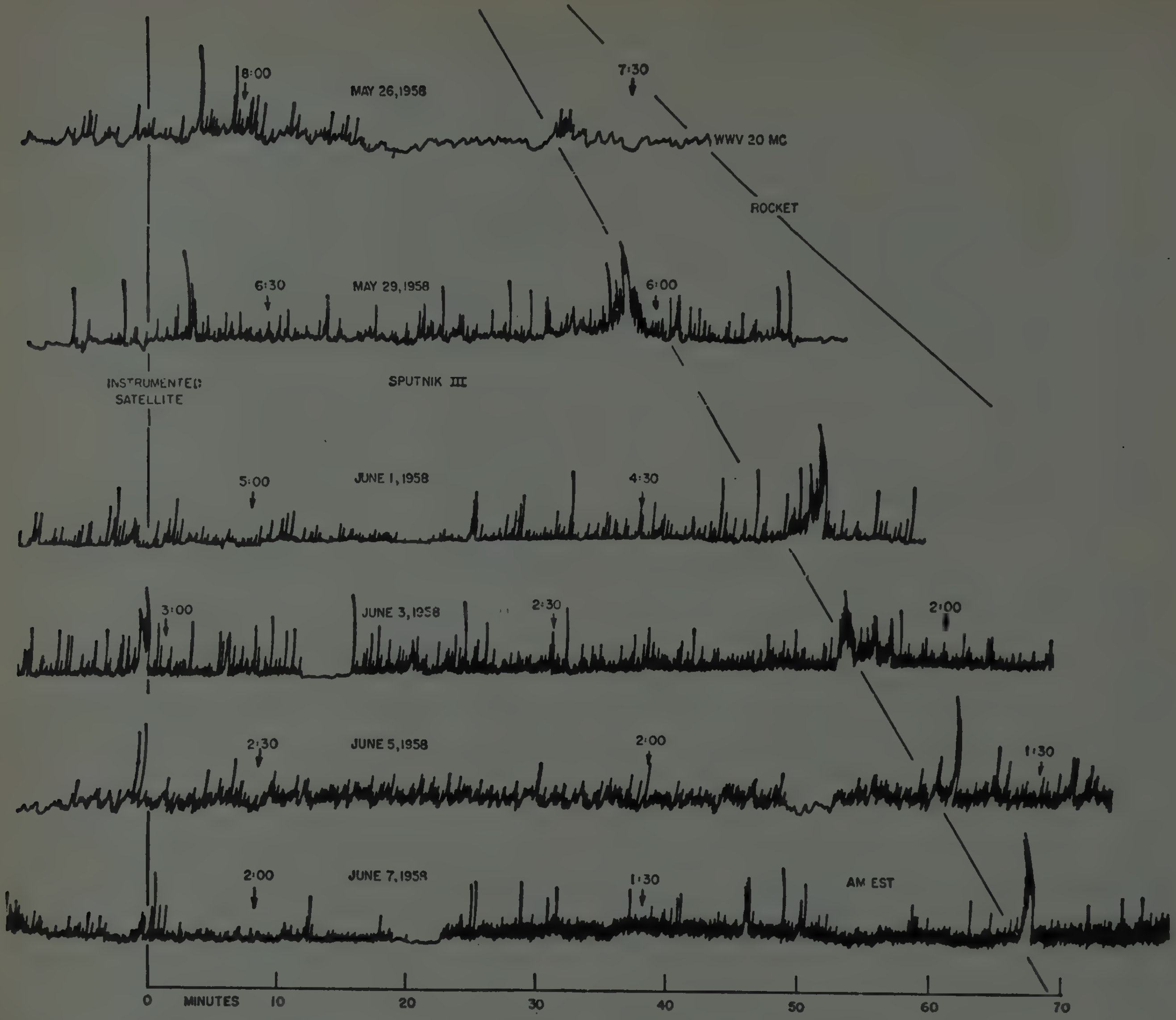

Fig. 2—Records of WWV 20-mc signals with relation to passes of Sputnik III. A large WWV burst is seen to move further ahead of the instrumented satellite on succeeding days. A line drawn through this series of bursts and extrapolated back intersects the vertical line for the instrumented satellite, and also a line for the rocket carrier, on May 15, 1958, the launching date, suggesting that the series of bursts may have been due to one of the other Sputnik III objects (such as the nose cone or side shields).

of artificial earth satellites, as described in the previous sections, suggest ionization induced by satellites. On many occasions, a much shorter period, of the order of 1 minute, was observed for groups of signal bursts occurring at the time of near approach of a satellite. In a typical example, a well-defined group of 10 to 15 bursts, each of 5- to 10-seconds duration, was observed to occur close to the time of near approach of Sputnik II. The spacing between bursts was remarkably constant, averaging 59 seconds with a mean deviation of 5 seconds. The amplitude of the bursts was in nearly all cases greater than that of bursts observed before or after the group. Furthermore, the largest bursts occurred near the center of the group. Between the bursts, the signal level was substantially zero so that the group had the appearance of a series of regularly-spaced spikes decreasing in amplitude from the middle to both the beginning and end of the group. It appears probable that these burst groups were related to Sputnik II. The fact that the tumbling period of the satellite was about 1 minute also favors this interpretation. However, the periodicity could have been related to the charge-discharge time of the satellite (see the section on Ionization Mechanisms) or to the spacing of relatively dense ionized bands or zones in the Van Allen belts or other regions through which the satellite was passing.

SYMMETRICAL BURST SEQUENCES

Groups of signal bursts differing from those described in the previous section have also been observed at The Ohio State University Radio Observatory. These burst groups are characterized by an even more symmetrical series of bursts with many of the bursts lasting about 1 minute. An example of such a symmetrical burst sequence is shown by the record of Fig. 3. The center trace is a record of the WWV 20-mc signal strength as a function of time for the morning of October 8, 1958. An Esterline-Augus recorder was used with time increasing from right to left.

A remarkably symmetrical series of bursts is evident with the center point at 4:22.8 A.M. EST. At this moment, the transmitterless Sputnik III rocket carrier was close to the north magnetic pole. It was also at its point of nearest approach to Columbus and was passing through the center of the beam of the receiving antenna. The geometry of the pass as viewed from the vicinity of the north pole is indicated by the sketch at the top of Fig. 3. The path of the Sputnik III rocket from west to east (across Hudson Bay) is shown by the track running from right to left with Columbus 1750 miles to the south (or above in the sketch). The curve marked 60° is the 60th parallel of latitude. Columbus is at 40° north latitude.

The scale along the path indicates the position of the satellite as a function of time. The scale of this sketch was chosen to correspond to the time scale of the chart record of the WWV bursts in the center trace. Thus, at about 4:15, when the satellite was approaching from the west, there was a large short-duration WWV burst. As the satellite approached closer, a series of more sustained bursts occurred which reached a maximum amplitude at about 4:23 when the satellite was in the center of the receiving antenna beam and close to its point of nearest approach. As the satellite traveled further east (to the left), the bursts decreased in amplitude until at about 4:31 when there was another large short-duration burst. Lines with arrows extending downward from the satellite path indicate the events occurring at the corresponding times on the WWV record. The record would suggest that the observed signals were in this case due to reflections of WWV signals from ionization induced by the Sputnik III rocket.

The receiving antenna used during these observations was a three-element horizontally-polarized beam antenna directed about 10° east of north. This direction was chosen as giving a minimum of response to WWV under ordinary conditions. The antenna power pattern is indicated in the sketch at the top of Fig. 3. The recorded response should be a function not only of this pattern, but also of the distance of the satellite path from Columbus.

In order, however, to obtain a calculated response curve which corresponds to the envelope drawn over the bursts at the bottom of Fig. 3, as shown by the dashed line, an additional factor is required which suggests a considerable aspect sensitivity to the response. That is, the strength of the signal bursts appears to fall off rapidly as the angle between the radius vector and satellite path becomes less than 90°, which is the condition for a specular reflection from an ionized column along the satellite path. The brief spikes of WWV signal observed at about 4:23 may also be momentary enhancements which occurred when the reflection was specular, and, if this is the case, the aspect sensitivity is even greater than suggested above since the envelope in Fig. 3 was not drawn over these spikes.

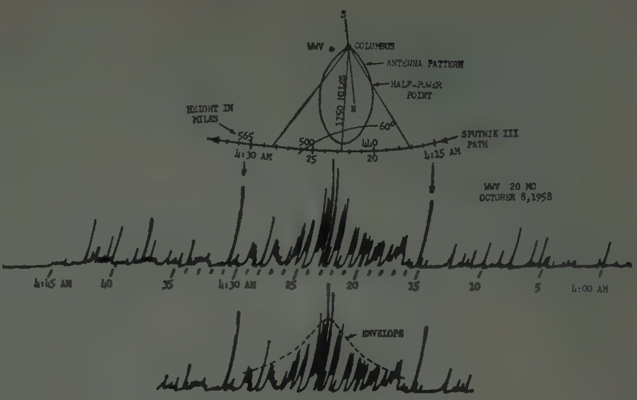

Fig. 3—Record of the WWV 20-mc signal obtained on October 8, 1958, during the passage of the Sputnik III rocket through the antenna beam (center trace). The sketch at the top of the figure shows the pass geometry as viewed from the vicinity of the north pole. The record of the WWV signal is reproduced again at the bottom of the figure with an envelope (dashed line) drawn over the burst pattern.

Magnetic sound tape recordings made of the 20-mc signals at the time of the pass reveal the characteristic WWV signal in many of the bursts. In the more sustained bursts, a rapid flutter fading is evident, and in some of the strongest bursts near 4:23, the fading is so severe as to mask the WWV signal characteristics in a very hashy type of random noise. High-speed paper tape recordings of the signals from the magnetic sound tape show a very definite amplitude modulation (about 50 per cent) of the reflected WWV signal with a period of about 9 seconds. This period is the same as the tumbling period of the satellite which suggests that the ionization produced by the satellite is a function of the physical cross section of the satellite normal to its path.

It may be significant that the Sputnik III rocket passed north of the center of the auroral belt at about 4:17 and south again at about 4:28, so that the satellite was north of the center of the auroral belt and near the north magnetic pole at the time the principal bursts were observed. Hence, it could possibly be that each burst marks an encounter of the satellite with denser bands of ionized particles situated along lines of the earth's magnetic field.

A calculation of the radar cross section for this symmetrical burst sequence indicates a value of several square kilometers. Since this area exceeds the physical cross section of the rocket (about 100 square meters) by many orders of magnitude, the implication is that large amounts of ionization were induced by the satellite.

The symmetry of the bursts in the record of Fig. 3 is summarized in Table I. Comparing the burst pairs 33′, 22′, and 11′ we should note that they are symmetrical in time with respect to the center of the sequence to within about 0.2 minute. Another striking feature of the symmetry is that the bursts at about 4:16 and 4:19 are nearly mirror images of the ones at 4:27 and 4:29.

TABLE I

Burst Number	Time of Burst	Time before (−) or after (+) nearest approach
3	4:07.8 A.M.	−15.0 minutes
2	4:15.1	− 7.7
1	4:16.5	− 6.3
0	4:22.8 (nearest approach)	0.0
1′	4:29.3	+ 6.5
2′	4:30.7	+ 7.9
3′	4:37.8	+15.0

JULY 23, 1958, OBSERVATIONS

As another example of WWV signal bursts associated with a satellite's orbital period, the entire record of the WWV 20-mc signals received on the morning of July 23, 1958, between 12:55 and 8:00 A.M. EST is presented in Fig. 4. This figure is a photograph of the original Esterline-Angus recording with time increasing from right to left. The record has been cut into four equal parts with the center of each section coinciding with the time of the near approach of the Sputnik III instrumented satellite (1958 Delta 2). The vertical line at the center of the figure indicates the approximate time of near approach and the slant line the approximate time of the 40th parallel crossing of the satellite (traveling NNW to SSE). The indications on the record are due to WWV and not to the Sputnik III transmitter (20.005 mc), since, as indicated by the top trace, the deflection dropped to zero level during the WWV off period between 1:45 and 1:49 A.M. even though the satellite was at its point of nearest approach during this interval and readily received on its frequency.

The traces in Fig. 4 are noteworthy since, on four successive passes of the Sputnik III instrumented satellite, there are significant deflections occurring close to the time of the near approach of the satellite to Columbus. Also, there are no deflections of much significance at times other than these.

The geometry of the passes of Fig. 4 was as follows. On the first pass (1:47 A.M.), the satellite (traveling NW to SE) made its near approach over northeastern Canada to the northeast of Columbus. On the second pass, the satellite traveling NNW to SSE made its near approach over the northeastern U. S. to the northeast of Columbus. On the third pass, the near approach of the satellite was to the southwest of Columbus, while on the fourth pass, the near approach was also to the southwest of Columbus but further in this direction.

DOPPLER EFFECTS

In some CW-reflection observations at The Ohio State University Radio Observatory, it was noted that the correlation of received WWV bursts with artificial satellite passes was improved when the receiver was slightly detuned from the WWV carrier frequency. The record of Fig. 4, discussed in the preceding section, is an example of this effect, the receiver of 5-kc bandwidth having been tuned so that the center of its response was about 2 kc below 20 mc. Subsequent observations with a swept-frequency receiver also resulted in excellent correlations between satellite passes and bursts of WWV signal shifted almost simultaneously to frequencies both above and below the carrier frequency. The observed frequency shift involved at 15 mc was characteristically at least 5 kc. This frequency shift is believed to be a satellite-related Doppler effect. As observed on a loudspeaker, the bursts have no readily distinguished audible tone, but appear to have the characteristics of pure noise.

An example of a series of such Doppler bursts at the time of a satellite pass is shown in Fig. 5. Here the swept-frequency film obtained on 15 mc [Fig. 5(c)] and the upper side-frequency record obtained on 10 mc [Fig. 5(b)] during a pass of the Sputnik III instrumented satellite about 5:55 A.M. (EST) on April 15, 1959 are presented.

The geometry of the pass is indicated in Fig. 5(a) as viewed from a point near the north pole. The locations of Columbus, Ohio and WWV (point of origin of signal) are at the top of the figure on or close to the 40th parallel of latitude. The parallels for 50° and 60° are also indicated. The path of the Sputnik III instrumented satellite is shown along the bottom of the sketch with the satellite direction of travel from right to left (west to east) with times (EST) for the satellite indicated along the path.

The WWV signal received in the frequency band from about 10.001 to 10.003 mc is shown in Fig. 5(b). Time increases from right to left as indicated. Just after Sputnik III had entered the auroral belt, about 5:50 A.M., there was a marked increase in signal reaching its maximum about 5:51 A.M., or one minute before the point of nearest approach to Columbus. Another increase in signal occurred between 5:55 and 5:57 A.M. The recordings of Fig. 5(b) were made using a horizontal half-wave dipole receiving antenna oriented east-west.

Fig. 5(c) shows the simultaneous recording on 35-mm film of the output of a swept-frequency receiver (center frequency 15 mc) between 5:49 and 6:00 A.M. The receiver was swept about twice per second from a frequency of 15 mc minus 4.3 kc (top of the film) to 15 mc plus 4.3 kc (bottom). The receiver bandwidth was about 1 kc. The heavy central trace is the 15-mc carrier of WWV (Washington, D. C.) which was off the air just

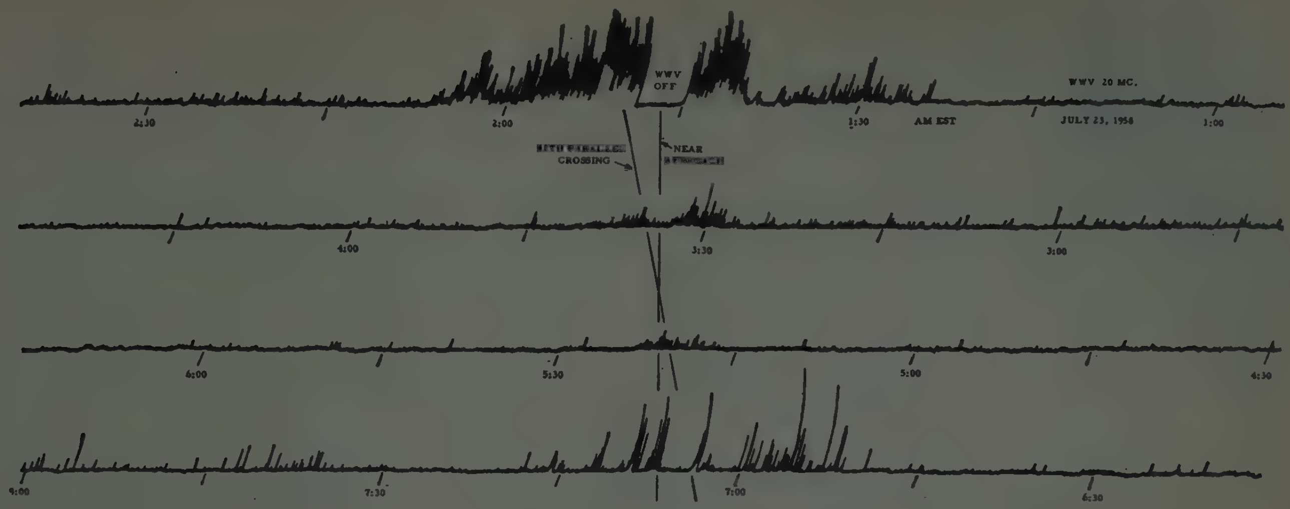

Fig. 4—Records of WWV 20-mc signals recorded between 12:55 and 8:00 A.M. EST on July 23, 1958, showing significant deflections at the times of the near approach of the Sputnik III instrumented satellite on four successive passes.

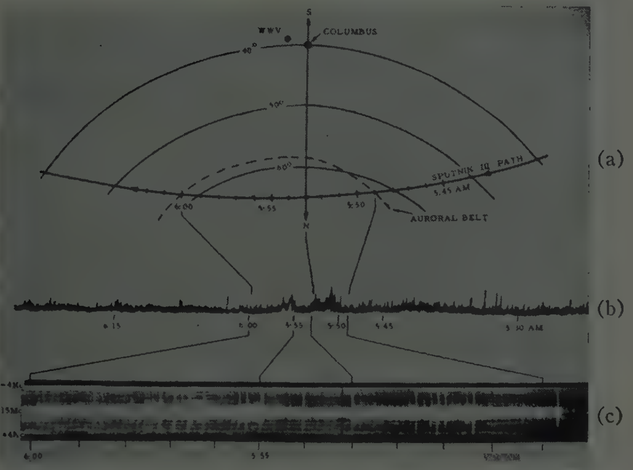

Fig. 5—Geometry (a) of path of the Sputnik III instrumented satellite on its 5:55 A.M. pass of April 15, 1959, with 10-mc WWV signal (b) recorded in the upper side-frequency band between 10.001 and 10.003 mc and the 15-mc WWV signal (c) recorded over the frequency range between 14.996 to 15.004 mc. Bursts of Doppler-shifted WWV 15-mc signal appear almost simultaneously in both sidebands (c) with intensity maxima near 5:51 and 5:57 A.M. corresponding to peaks of 10-mc signals in the upper side frequency at the same times (b).

prior to 5:49 A.M. The 600- and 440-cps tone modulations of WWV, particularly in the upper sideband, are discernible between 5:50 and 5:53 and again between 5:55 and 5:58 A.M. The absence of sideband modulation is also clearly indicated between 5:53 and 5:55 A.M. Of particular interest on this film are the bursts which appear as slightly curved, almost vertical lines. In many cases, these lines extend from the upper- to the lower-frequency limits of the film and perhaps farther. These bursts are especially strong around 5:51 and 5:57 A.M. These times correspond approximately to the times of the signal peaks observed on 10 mc [Fig. 5(b)]. These bursts appear to be Doppler-shifted components of WWV signal which have been deviated by as much as 4 kc or more.

The receiving antenna for the 15-mc observations was a horizontally-polarized directional antenna of the corner reflector type with a physical aperture of 1360 square feet. This directional antenna, with a horizontal half-power beamwidth of about 50°, rotated in azimuth about 6 times per minute. At the instant it passed a fixed reference direction, a fiducial mark was impressed on the film. By interpolation, the direction of arrival of a burst of signal on the film can be determined with an accuracy of about ±30°. Making use of this direction information, we found the azimuthal direction of signal arrival of the bursts in the upper sideband in Fig. 5(c) to be NNW at first, changing gradually to NNE, while in the lower sideband, the direction of the bursts was generally south. Doppler shifts into the upper sideband indicate approach of an object, while shifts into the lower sideband indicate recession. Hence, these observations suggest that the Doppler-shifted signal bursts were produced by reflection of the WWV signal from ionized clouds or streams traveling from the north to the south. The Doppler shifts of at least 4 kc at 15 mc indicate cloud velocities of at least 40 km/second. Since the clouds appear to come from the satellite's general direction, the implication is that they are satellite induced.

In addition to the effects described above, several other Doppler effects have been noted on occasion. One is a coherent Doppler signal changing smoothly in pitch approximately as would be expected for signals reflected from an ionized cloud moving with the satellite. This, however, is a rare effect. A second more commonly observed effect is a quasi-coherent Doppler signal of very irregular characteristics. Such Dopplers have a warbling sound and consist of audio components covering many hundreds of cycles. Such Dopplers may come in groups at the time of a satellite pass and suggest the passage of a number of ionized clouds which may be satellite induced. A third type of Doppler phenomenon occasionally noted is observed as a flutter fading of WWV which begins at a rapid rate, goes through zero beat, and again increases to a rapid rate in a manner analogous to the airplane flutter commonly observed on local TV and FM stations. This effect has been noted at the near approach of artificial earth satellites and is presumably satellite-related since the distance of WWV is too great for an airplane to produce the effect.

Another occasionally observed effect involves a constant pitched tone of the order of 350 cps which is heard superposed on the WWV signal at the time of the pass of a satellite. A frequency of 350 cps corresponds to the gyro-frequency for protons in a magnetic field of 0.23 gauss, a value which occurs in the earth's magnetic field at high altitudes in the lower latitudes. Hence, a proton interaction mechanism is a possible explanation. This constant-pitched tone effect was first reported by Roberts, Kirchner, and Bray[6]. They also were the first to report observing Doppler signals similar to the second type described in the preceding paragraph.

Ionization Mechanisms

The velocity of artificial earth satellites is typically of the order of 7 km/second. Particles initially at rest can be driven ahead of the satellite by an elastic collision at twice this velocity, or about 14 km/second.[5] This is only a fraction of the 40-km/second velocity observed in the clouds discussed in the preceding section. If these clouds are actually satellite induced, as seems to be indicated, then some other acceleration process must be present or else the effect is due simply to an encounter of the satellite with fast corpuscular streams, the particles being driven ahead, deviated, or scattered by the satellite. The noncoherent nature of the reflected signal may be due to such things as turbulence or a random distribution of velocities in the deflected clouds or streams. Assuming elastic collisions and an exactly head-on geometry of encounter, the resulting particle velocity is equal to the incident stream velocity plus twice the satellite velocity.

If the satellite acquires an electric charge, its collision cross section for ionized particles will be increased and the number of particles swept ahead will be larger. Jastrow has indicated that satellites may charge up to potentials as high as 1000 volts.[7] The sweep-up effect of the satellite will also tend to leave a deficiency of particles immediately behind it. For example, if the satellite has a negative charge it will sweep electrons ahead while trailing behind will be an electron hole. This hole in the wake of the satellite constitutes a discontinuity in the medium and may account for significant amounts of scattered signal.

Conclusion

In conclusion, it may be mentioned that many of the best correlations between WWV signal effects and passes of artificial earth satellites have been observed to occur a day or so after large solar flares. At such times the density at high altitudes, particularly in the auroral regions, is increased by the influx of solar particles and it is reasonable to expect that an artificial earth satellite passing through such regions may produce enhanced ionization effects.

Acknowledgment

The assistance of D. J. Scheer in obtaining the 10-mc record of Fig. 5(b) and of R. T. Nash in the construction of the rotating corner reflector is gratefully acknowledged.

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[6] C. R. Roberts, P. H. Kirchner, and D. W. Bray, "Radio detection of silent satellites," *QST*, vol. 43, pp. 34–35; August, 1959.

[7] R. Jastrow, "Artificial satellites and the earth's atmosphere," *Scientific American*, vol. 201, pp. 37–43; August, 1959.

Interplanetary Telemetering*

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Summary—The use of telemetry systems over interplanetary distances generates performance requirements more severe than ever before experienced. Full exploitation of the foreseeable state-of-the-art and high over-all operational efficiencies will be essential in fulfilling the requirements of these rigorous standards of performance.

The problem areas associated with interplanetary telemetry and an approach to the selection of system operating parameters are discussed. The analysis is based on the dividing of operational efficiency into two categories: information efficiency and physical system efficiency. The performance of a typical interplanetary system is analyzed to exemplify the importance of a number of the operating parameters which affect system performance.

INTRODUCTION

THE USE of telemetry systems over interplanetary distances generates performance requirements never before experienced. These rigorous performance requirements will demand full exploitation of the foreseeable state-of-the-art and high over-all operational efficiencies in order to achieve successful interplanetary telemetering or communication. This over-all operational system efficiency may be divided into two categories, one that may be called physical system efficiency, and the other, information efficiency.

Physical system efficiency is determined by such tangible factors as system power and power-conversion efficiency; antenna gains; free space, atmospheric and other signal attenuations and noise at a given radio frequency; receiver and antenna noise levels; and other physical factors.

It is beyond the scope of this report to analyze in great detail the information theory aspect of communication. However, it is a very important consideration and the information efficiency of an interplanetary telemetry system will have a direct bearing on system power requirements. For this reason, a brief discussion of information efficiency is presented, even though this report deals mainly with the physical factors affecting interplanetary telemetry.

INFORMATION EFFICIENCY

Very elementary information theory indicates that systems for other interplanetary communications must employ the same techniques that will be used for telemetry over the same distances. There will be little to distinguish between the signal parameters and much of the apparatus of interplanetary communication and interplanetary telemetry, because of the necessity for both systems to convey the maximum intelligence with the least bandwidth and power demands.

Coding is the essence of applied information theory. When the rate of transmission is predetermined, theory can show the most economic combination of bandwidth and power. However, the current state-of-the-art, limited primarily by technical factors of the coding and decoding devices, is such that no practical coding system has been developed which fully realizes the efficiencies possible in theory. In telemetry, pulse code modulation (PCM) is the practical development which has coincided most closely with theoretical studies of information theory and is one of the most efficient systems in operational use. The bandwidth and information are both linearly proportional to the number of digits, with fixed pulse width, so that the information capacity of a PCM system is proportional to the bandwidth.[1] This is in accordance with the results promised by ideal coding. Conventional PCM, on the other hand, does not make each received information component dependent upon the resultant of a number of signal elements, and in the absence of such an averaging process it cannot work down to a 1:1 ratio of signal power to average noise power. Also, the quantizing and reconstruction of the signal produces some differences between the received information and the original; this difference is called quantizing noise, a function of the coding system itself. Numerous other considerations could be mentioned; however, the above illustrates that even PCM does not represent the optimum coding scheme based on performance theoretically possible.

The preceding discussion has shown that current coding systems are not ideal, and thus new coding systems, approaching more closely to the ideal and therefore more efficient, are bound to be developed to reduce the other system requirements; this is an essential step for reliable interplanetary telemetry and/or communications. It is possible that these new coding techniques will bear little or no resemblence to the currently known techniques. The orthogonal modulation scheme behaves much like the ideal system described by theory and is gaining increased acceptance because of the current interest in high information efficiency.[2] A practical approach to an orthogonal system is the PCM system employing a redundant code, such as the Reed-Muller. The development of several such systems is known to be currently underway.[3,4]

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* Original manuscript received by the IRE, December 2, 1959.

† Research Div., Radiation, Inc., Orlando, Fla.

[1] W. Jackson, "Communication Theory," Academic Press, New York, N. Y.; 1953.

[2] R. W. Sanders, "Communication efficiency of space telemetry systems," *Proc. Natl. Telemetering Conf.*, Denver, Colo., pp. 4–13; May, 1959.

[3] H. N. Putschi and E. Niemann, Jr., "The PCM-PS telemetry system," *Proc. Natl. Telemetering Conf.*, Denver, Colo., pp. 143–150; May, 1959.

[4] R. W. Sanders, "Digilock telemetry system," *Proc. Natl. Symp. on Space Electronics and Telemetry*, San Francisco, Calif., sec. 6.3, pp. 1–10; September, 1959.

Another approach to the problem of increasing the information efficiency of the over-all telemetry system is that of performing in-flight data reduction or processing through the use of statistical techniques.[5] The most useful application of this approach involves the handling of vibration data. In this case, an in-flight real-time spectrum analysis of the data can reduce the modulation rate by 1000:1, since only the characteristics of the spectrum are transmitted.

Table I[6] contains a comparison of four characteristics—threshold carrier signal strength, RF power, information efficiency,[7] and RF bandwidth—for sixteen possible telemetry systems. This table is based on an individual channel output signal-to-noise ratio at threshold of 100 and a total information bandwidth of 1000 cps. The development of the equations used in determining values for the table has been concisely presented.[6]

TABLE I

Comparison of Telemetry Systems

Type	Threshold Carrier Signal Strength	RF Power (Relative to PPM/AM)	Information Efficiency	RF Bandwidth (Kilocycles)
1. PPM/AM	200	1	0.17	76
2. PCM/FM	260	1.7	0.24	18
3. PCM/PM	280	2.0	0.21	20
4. PCM/AM	370	3.4	0.21	18
5. PAM/FM	580	8.3	0.050	85
6. AM/FM	610	9.3	0.045	93
7. PDM/FM	610	9.3	0.045	92
8. PDM/PM	660	11	0.036	110
9. FM/FM	740	14	0.030	140
10. AM/PM	770	15	0.028	150
11. PAM/PM	780	15	0.028	150
12. PDM/AM	790	16	0.035	94
13. FM/AM	830	17	0.055	50
14. FM/PM	860	18	0.023	185
15. PAM/PM	3150	250	0.073	18
16. AM/AM	9600	2300	0.24	9.5

Physical System Efficiency

Establishment of the operating parameters of an interplanetary telemetry system is a rather involved process of trade-offs and compromises. This is because of the many interdependent factors and the need for system optimization. Computer techniques have been logically employed as an aid in determining the optimum system for a particular application.[8]

The fundamental factors limiting the performance[9] of an interplanetary telemetry system may be considered to be system reliability, information capacity, and power consumption for a given physical configuration. Since much has been written regarding component and system reliability it will not be treated in this paper. For the purposes of this discussion, it is assumed that reliability will not limit system performance. Thus, the basic design requirement becomes one of maximizing system operational efficiency. As shown above, operational efficiency may be divided into two categories, physical system efficiency and information efficiency.

Physical system efficiency is determined by the physical factors influencing system performance; in fact, all factors other than information efficiency fall into the category of physical system efficiency. These factors include:

1) Power conversion efficiency.
2) Transmission losses (free-space attenuation, oxygen and water-vapor absorption, line losses, Faraday effect, etc.).
3) External noise (galactic background noise, cosmic and solar noise, man-made noise, etc.).
4) Internal noise (receiver and antenna noise).

The selection of transmission frequency is influenced to some degree by all of the above factors. A transmission "window" in the RF spectrum exists from about 1 to 10 kmc. This is illustrated in Fig. 1 which shows ap-

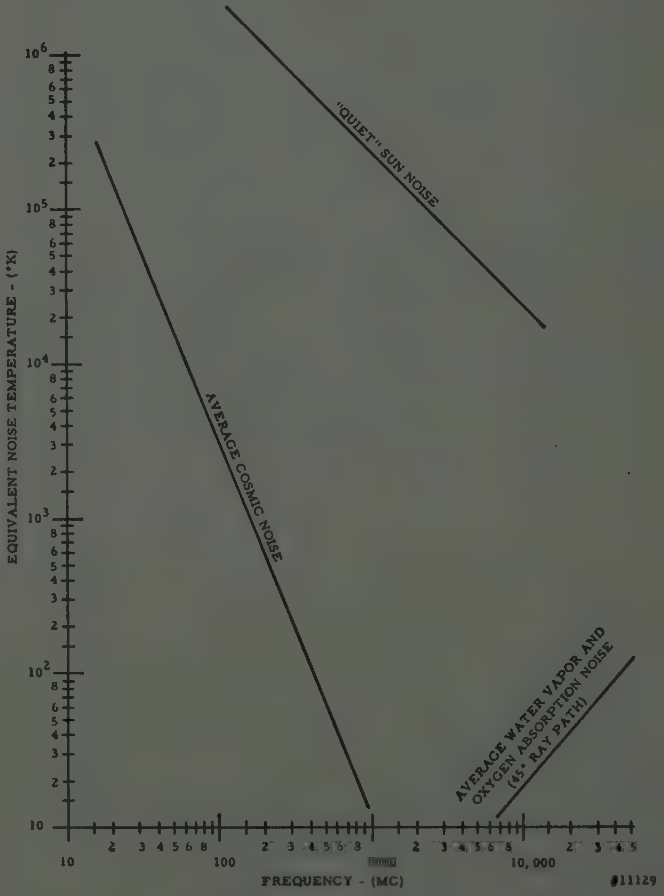

Fig. 1—Ambient space noise vs frequency.

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[5] L. G. Zukerman and I. Ross, "Some coding concepts to conserve bandwidth," *Proc. Natl. Telemetering Conf.*, Denver, Colo., pp. 138–142; May, 1959.

[6] M. H. Nichols and L. L. Rauch, "Radio Telemetry," John Wiley and Sons, Inc., New York, N. Y.; 1956.

[7] Defined here as the ratio of the information capacity of the output signal channel to the information capacity of the modulated signal.

[8] D. R. H. White, "Design and evaluation of space communications systems using computer simulation techniques," *Proc. Natl. Symp. on Space Electronics and Telemetry*, San Francisco, Calif., sec. 1.1, pp. 1–35; September 20–30, 1959.

[9] Performance is defined here as the total information-transmission capability of the system, a function of both information rate and transmission time.

proximate average cosmic and solar noise, and water vapor and oxygen absorption noise as a function of frequency. Certain advantages in operating at higher frequencies are apparent, *i.e.*, reduced cosmic and solar noise, increased antenna gains (for a given size), etc.; however, present receiver and transmitter state-of-the-art necessitates operation near the lower edge of the "window." The *S*-band radio telemetry allocation is very well suited for interplanetary use and much effort is being made to develop new telemetry receivers and transmitters capable of satisfactory and efficient performance at 2250 mc. Sources of noise external to the receiver input, at this frequency, are such that system sensitivity will be normally limited by the receiver input noise, except in the case of maser or parametric amplifiers. Excluding solar noise, the noise of these amplifiers is of the same order of magnitude as the sum of the external noises. However, under certain conditions, a receiving system incorporating a maser amplifier will be limited in threshold sensitivity primarily by external noise.[10]

The minimum level of transmission power required for a specific application is difficult to calculate because of the numerous factors affecting signal propagation. Since little inclusive empirical data are available it is necessary to include a rather large fading margin as a safety factor. Fig. 2 illustrates the free space attenuation between Earth and the near planets. Free-space attenuation is the single factor having the greatest influence on transmission power requirements. Although this is a parameter fixed by the maximum range of each specific application, the more important of the other factors determining transmission power are controllable parameters of the system.

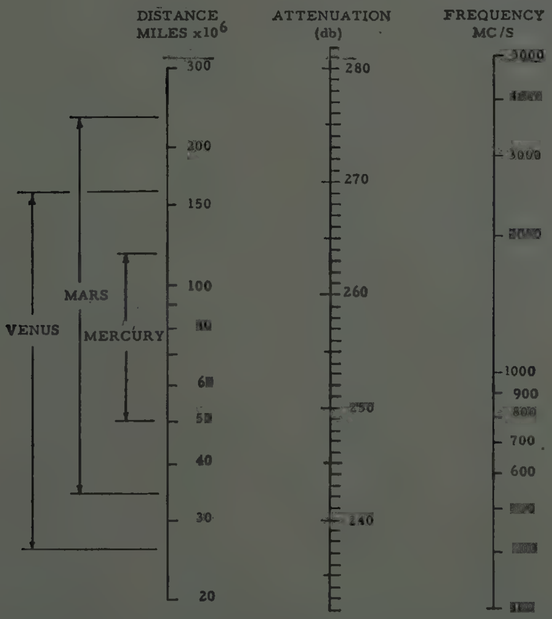

Fig. 2—Nomograph for determination of free-space path attenuation between isotropic antennas.

The bandwidth of the system greatly influences transmission power sensitivity and must be carefully selected because of its wide latitude. A range of 100,000:1 in bandwidth (representing a 50-db power differential) is typical of current communication systems. Because of other existing limitations, it is expected that the desired bandwidth of early interplanetary telemetry systems will be compromised to a value of not more than approximately 500 cps.

Antenna gains will be limited chiefly by the physical size which can be tolerated and the beamwidth limitations imposed by antenna-system tracking capability. The limitations of the vehicle-mounted antenna will be, of course, more severe than those of the ground-based antenna. However, the development of fold-out vehicle antennas and stabilized platforms will allow a substantial improvement over the relatively small antennas which are currently practical. Even a parabolic vehicle antenna of five feet in diameter and having a beamwidth of 6° (at 2250 mc), which now may be feasible, would probably have to be of the fold-out type.[11] Probably the most important limitation restricting the size of a parabolic ground-based antenna is the tolerance to which the reflector can be constructed. Target acquisition becomes increasingly difficult as the beamwidth decreases; but with a known trajectory, acquisition and tracking should be practical with beamwidths of less than 1°.

Receiver detection threshold is a factor which varies within rather narrow limits (compared to the effects of the previously discussed system parameters) and is a function of the type of carrier modulation and detector. Correlation-detection techniques have been applied to FM receivers as phase-locked detectors with threshold improvements in the order of 6 db. An analysis of this aspect of the system logically falls into the category of information theory.

The system power requirement (and thus power conversion efficiency) is another essential consideration. This will determine the capacity of a system which can operate from a given energy source, or conversely, the energy required for a particular system and operating time. Power conversion efficiency of a transmitter can vary from 5 to almost 50 per cent, depending upon the modulation requirements, whether it employs tubes or solid-state components, operating frequency, power output, etc. Over-all telemetry system power requirements are such that the rest of the system may require as much, or more, power than transmitters operating at levels below about 50 watts. Solid-state systems will be very valuable in minimizing power requirements for interplanetary missions. It is expected that even the first interplanetary systems will be almost entirely transistorized. In the future, highly efficient solar cells may completely remove the current power requirement restrictions.

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[10] R. Hansen, "Low noise antennas," *Microwave J.*, vol. 2, pp. 19–24; June, 1959.

[11] H. Scharla-Nielsen, "Space ship telemetering," presented at Natl. Symp. on Telemetering; April, 1957.

It can be seen from the preceding discussion that physical system efficiency or over-all telemetry system performance, exclusive of coding or modulation considerations, is influenced by a large number of factors of varying importance. Each one of these factors must be carefully analyzed with respect to the total system and mission to enable design parameters to be established which will yield the most efficient practical interplanetary telemetry system.

Factors Influencing System Performance

Transmission losses are generally thought of as including free-space attenuation, water-vapor and oxygen absorption, etc. It is frequently more convenient to consider many of these factors in terms of equivalent noise temperature, rather than in terms of power loss. This allows transmission loss factors and noise sources (receiver, cosmic, solar, etc.) to be treated in the same manner. This approach is especially valuable when dealing with maser and parametric amplifiers. The performance of these devices is best measured by techniques which yield performance figures in terms of equivalent input noise temperature. Also, noise figures of less than one decibel can be obtained.[12] These are more meaningful when expressed as noise temperatures.

Fig. 3 illustrates the effect of receiver, or receiving-system, effective input noise power (expressed as noise figure or effective noise temperature) on noise level over a range of receiver bandwidths. The basis for this information is shown by the following analysis.

The noise figure of a network is defined as

$$F = \frac{\frac{S}{N}\ \text{(input)}}{\frac{S}{N}\ \text{(output)}} \tag{1}$$

where S = signal power and N = noise power.

Theoretical ambient noise power is defined as KT_0B, where K = Boltzmann's constant (1.38×10^{-23} joules/°K), $T_0 = 290°\text{K}$, and B = bandwidth. By defining the effective internally generated noise power, referred to the network input, as P_n, the equation for noise figure[13] can be expressed as

$$F = \frac{\dfrac{S}{KT_0B}}{\dfrac{GS}{GKT_0B + GP_n}} = 1 + \frac{P_n}{KT_0B} \tag{2}$$

where G = network gain; or,

$$P_n = (F - 1)KT_0B. \tag{3}$$

Expressing P_n as KT_eB, where T_e is defined as the effective input noise temperature of the network (internally generated and referred to the input), (3) becomes

$$KT_eB = (F - 1)KT_0B,$$

or

$$T_e = (F - 1)T_0, \tag{4a}$$

also

$$F = 1 + \frac{T_e}{T_0}\cdot \tag{4b}$$

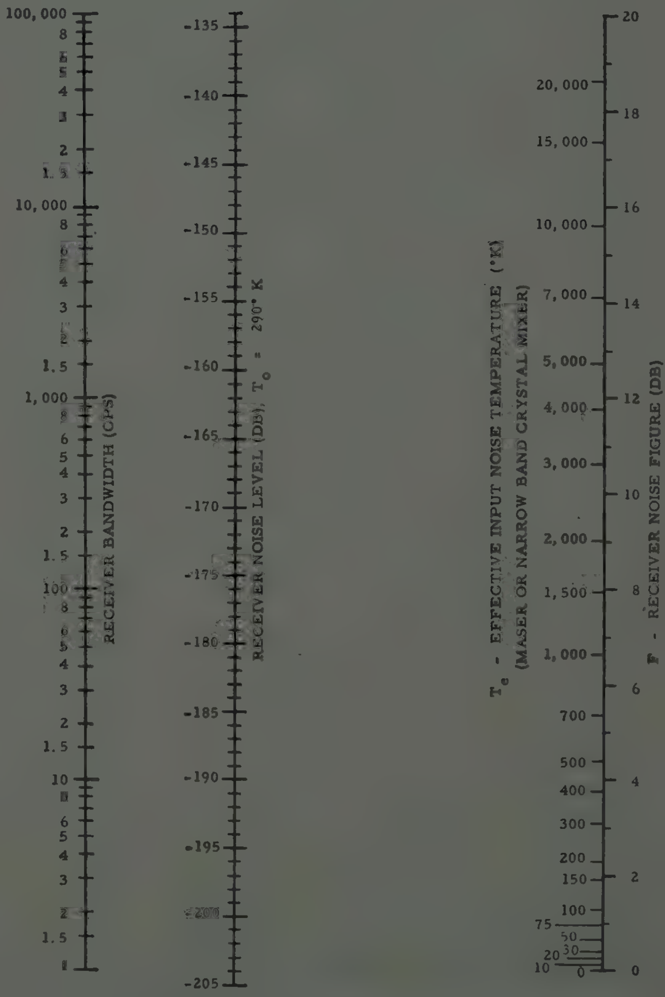

Fig. 3—Nomograph for determination of receiver noise level.

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[12] S. Perlman, *et al.*, "Concerning optimum frequencies for space vehicle communications," *Record of Natl. Symp. on Extended Range and Space Communication*, Washington, D. C., pp. 24–31; October, 1958.

[13] H. I. Ewen, "A thermodynamic analysis of maser systems," *Microwave J.*, vol. 2, pp. 41–46; March, 1959.

The total noise power, P, of a network is composed of the sum of the theoretical ambient noise power and the internally generated noise power. When these are added [(3) plus KT_0B], the following familiar expression is obtained:

$$P = (F - 1)KT_0B + KT_0B,$$

or

$$P = FKT_0B. \tag{5}$$

Substitution of (4b) in (5) for F illustrates that noise temperatures T_0 and T_e may be added to obtain the total noise power.

$$P = \left(1 + \frac{T_e}{T_0}\right)KT_0B,$$

or

$$P = (T_0 + T_e)KB. \tag{6}$$

It can be seen from (4)–(6) that in a noiseless network where $F=0$ db, or 1, $T_e=0$ also, so that $P=KT_0B$.

Fig. 1 shows that water vapor and oxygen absorption noise is not a significant factor below 10 kmc when the ray path is 45°, or more, from the horizon. As the transmission path approaches the horizon attenuation can increase by a factor of two; however, at a frequency of 2250 mc, this attenuation is negligible. Fig. 1 also shows that cosmic noise is insignificant at 2250 mc. The equivalent noise from specific radio stars will be higher but the average value is considerably less than 10°K. It can be seen that the sun is a strong noise source throughout the entire RF "window." Even though solar noise decreases at higher frequencies, the current state-of-the-art is such that an attempt to operate at the apparently ideal frequency of 8–10 kmc would not be practical. Also, the sun as a noise source can be avoided since its angular width is small. At 60 mc this angle may approach 1° and at 30 kmc the angle is approximately 0.5°. Thus, it can be seen that the radio frequency of 2250 mc represents a logical compromise between present communication techniques and galactic noise levels.

Free space attenuation between Earth and Venus, Mars, and Mercury can be determined from Fig. 2. As an example, the free-space attenuation for a typical interplanetary distance of 60 million miles (approximate median distance to Mars, Venus, and Mercury) is 260 db at 2250 mc.

With the successful application of maser and parametric amplifiers, the receiver is no longer the factor which limits system sensitivity, even at microwave frequencies. Antenna noise temperatures (largely ignored until recently) may be as high as 150°K and frequently limit system sensitivity when the above amplifiers are employed. Noise sources include the atmospheric, cosmic, and solar noise levels previously discussed. Additional noise results from the side and back lobes, as well as the main beam, "seeing" land and even sea areas as thermal reservoirs.[10] Also, losses in the antenna feed and reflector add to the noise temperature. Radomes represent another noise source, because of losses in the dielectric and sometimes high ambient temperatures.

Analysis of Transmission Power Requirements

Using the typical interplanetary mission of Venus probe as an example, RF power requirements have been calculated based upon the following parameters.

Distance (nominal) $=60\times10^6$ miles
Radio frequency $=2250$ mc
Transmitting antenna:
  Diameter $=5$ feet
  Gain $=29$ db
  Beamwidth $=6°$
Receiving antenna:
  Diameter $=100$ feet
  Gain $=54$ db
  Beamwidth $=0.3°$
  Noise temperature $(T_{LB}) = 55°)$ K[14]
Passive element losses $(L_0) = 1$ db (approximate).

A composite three-term expression[13] for the effective noise temperature of the system is given by:

$$T_s = \underset{1)}{[T_G + T_{BL}]} + \underset{2)}{[(L_0 - 1)T_0]} + \underset{3)}{[T_eL_0]}. \tag{7}$$

1) Temperature contributions from sources outside the signal path.
2) Radiation temperature of composite losses, amplified by the magnitude of the composite loss.
3) Effective temperature of the active circuits, amplified by the magnitude of the composite loss forward of these circuits. Where

T_G = galactic background radiation field, cosmic noise.

T_{BL} = produced as a result of the inefficiency of the antenna in directing its entire pattern into the forward direction (back lobes intercepting the earth as a thermal reservoir).

L_0 = composite losses of the passive hardware of the system.

T_e = effective temperature of the active circuits.

$T_0 = 290°$K.

Solving (7) yields a value of T_s, from which an overall system noise level can be calculated. Using the given system parameters and a nominal sky temperature of

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[14] Cold sky at S band.

10°K (excludes the sun and discrete radio stars, such as Centaurus, etc.), T_s becomes:

$$T_s = [10 + 55] + [(1.259 - 1)290] + [25 \times 1.259] = 172° \text{ K}.$$

Taking the necessary bandwidth (B) as 300 cps and solving (6) yields a value for P as follows:[15]

$$P = (290 + 182)(1.38 \times 10^{-23})(300) = -177 \text{ db}.$$

Fig. 3 can be used to obtain this value by our drawing a line between T_e of 172°K and B of 300 cps. This line intersects the receiver noise level scale at −177 db.

Required transmitter power may now be determined by the following procedure. A 20-db fading margin, or required SNR, is used to provide what should be an ample safety factor.

Receiving system noise level	−177 dbw
Required SNR	20 db
Receiving system sensitivity	−157 dbw
Receiving antenna gain	− 54 db
Path attenuation at 60×10^6 miles	260 db
Required radiated power	49 dbw
Transmitting antenna gain	− 29 db
Required transmitter power	20 dbw (100 watts)

This approach can be somewhat simplified through the use of Figs. 3 and 4. Fig. 4 also can be used to determine the gain of a parabolic antenna at various frequencies.

Basic transmission power required (Fig. 4) (with $D = 60 \times 10^6$ miles and $f = 2250$ mc)	56 dbw/cps bw
Correction for 300 cps bw	25 db
System noise factor (Fig. 3) (with $T_s = 172°$K)	2 db
Required SNR	20 db
Receiving antenna gain	−54 db
Required radiated power	49 dbw

CONCLUSIONS

The present state-of-the-art in radio telemetry is such that successful interplanetary telemetering can be achieved; however, to keep the transmission power requirements within reasonable limits, information rates will have to be low (possibly 500 cps) and the system will have to be carefully designed for maximum operational efficiency. It has been shown how the over-all operational system efficiency may be divided into two distinct categories, information efficiency and physical system efficiency, to facilitate analysis of system performance.

Applied information theory can show how the greatest information per unit bandwidth and power, or information efficiency, can best be obtained. Coding techniques are the products of applied information theory. Even the best of current coding schemes, and their variations, do not reflect the information efficiency possible in theory. Therefore, much is yet to be accomplished in this field and these results will appear as new coding schemes more closely following the level of efficiency promised by theoretical studies.

Factors influencing physical system efficiency are many and varied. The more important factors include: system power conversion efficiency, transmission losses, external noise, and internal noise. These factors, along with transmission frequency, are all to some degree interrelated. External noise sources (cosmic, solar, etc.) are distributed such that they limit the usable interplanetary transmission frequencies to a lower value of about 1 kmc. The upper frequency limit of about 10 kmc is determined by water vapor and oxygen absorption. Thus a transmission "window" exists between the approximate frequencies of 1 and 10 kmc.

The development of very low noise amplifiers, such as the maser and the parametric, represents a breakthrough of major significance to extended range communications. The receiver itself, previously the weakest link in receiving system sensitivity, has now become one of the strongest. Noise temperatures of 25°K (corresponding to a noise factor of substantially less than 1 db) are now being obtained with maser amplifiers, and future masers are expected to operate at noise tempera-

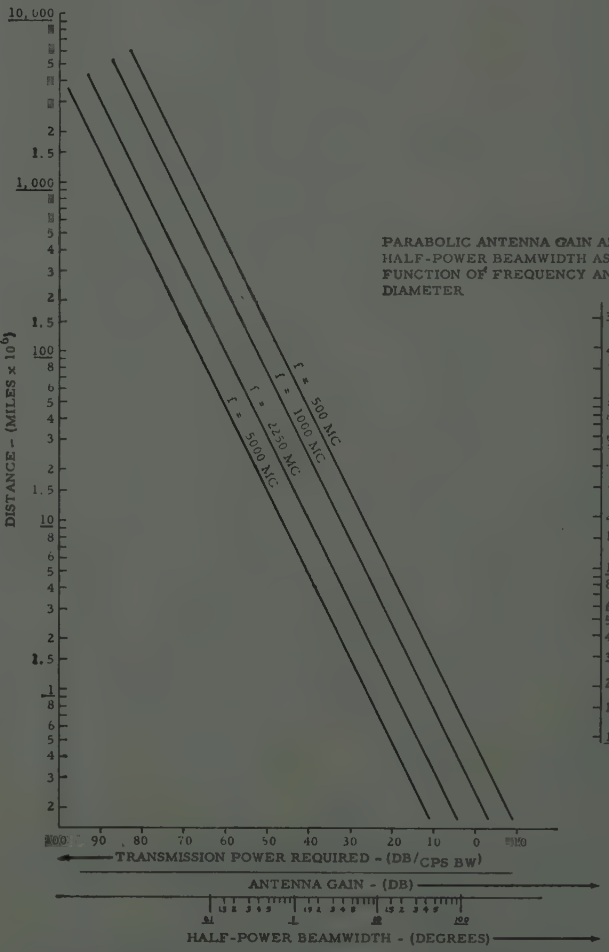

Fig. 4—Transmission power required per cps beamwidth at given frequency and distance. Isotropic antennas: $F = 0$ db; $S/N = 0$ db.

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[15] Since T_s is substituted for T_e, P becomes the total noise power of the system.

tures of 5°K. Parametric amplifiers are reported to be capable of noise temperatures of 75°K (approximately 1 db) at 1 kmc. Noise sources external to the receiver itself generally have now become the limiting factors of receiving system sensitivity. Thus, these external noise sources, including the antenna itself, are now being considered with an interest heretofore nonexistent (except possibly in the field of radio-astronomy).

The hypothetical system analyzed in the previous section is admittedly rather optimistic in some respects, especially in antenna characteristics. However, it is likely that steerable ground and vehicle antennas as described will be utilized in the near future. The system bandwidth of 300 cps is a limiting factor in information capacity, and it is expected that this situation will exist for some time until more efficient coding methods are put into practice and more efficient solar power sources are developed.

There can be no doubt, however, in spite of the problems involved, that reasonable amounts of intelligence can be telemetered back to Earth from the near planets by the use of techniques presently available. The approach to the problem presented in this report is by no means all-inclusive but it should give the reader a better understanding of this branch of communications for interplanetary applications.

Space Telemetry Systems*

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Summary—This paper presents some of the problems which are faced in the field of space telemetry, and gives a few examples of the approaches used in attempts to solve these problems to date.

The different types of information to be transmitted back to receiving stations on the earth from satellites and space probes are discussed. The telemetry systems (FM/FM, FM/AM, PCM/PM, etc.) which have been used in various experiments of this type are described, along with some typical performance figures for some past payloads.

The paper is concluded with a discussion of the problems to be faced in future experiments.

Introduction

THIS PAPER presents some of the problems which are faced in the field of space telemetry, and gives examples of the approaches which have been used to solve these problems. A summary of the characteristics of the telemetry systems used in the nonmilitary satellites and space probes is also given.

The launching and successful orbit of the first man-made earth satellite marked the beginning of a new era in the field of telemetry. For years man had used remotely located instruments to measure events and physical phenomena, first with "hard-wire" links from inaccessible or physically dangerous locations and later with the use of radio links. With the entry into space, man now faces a new field and new problems. This newly opened region, beyond the protective blanket of atmosphere, contains many unanswered questions. Measurements must be made of the space environment near the atmosphere to confirm or refute ideas based on extrapolated data and scientific conjecture. In addition, there is knowledge to be gained about the moon and the deep reaches of outer space.

Only after considerable knowledge has been gained about radiation levels, micrometeorite densities, temperature balance, and many other environmental factors will it be reasonably safe to send a man into this unexplored frontier. When manned space flight begins, telemetry will be required to bring back data on man's physiological and psychological conditions in his weightless environment.

Space Systems vs Conventional

In the testing of aircraft and in many ground-based systems, flexibility, ease of adjustment, ease of maintenance, and long life were factors of primary importance. With the advent of the missile age, size and weight of equipment took on new meaning and emphasis. It was found that in most missile operations, the life of the telemeter unit was limited to a single flight, and the addition of each pound of weight resulted in a serious degrading of the missile performance. Hence, a great amount of time and effort went into finding ways to improve the effective bandwidth utilization and reliability while reducing weight at the expense of long operating life and ease of operation.

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With the entry into the space age another re-evaluation of telemetry requirements must be made. Missile flights were usually limited to minutes, or possibly to a few hours for some air-breathing cruise-types, but satellites and space probes will have lifetimes of from weeks to years. Reliability and long life cannot be overemphasized because of the difficulty and great expense of each project. Power vs bandwidth problems become even more critical at the increased distances. Weight has increased in importance because of the penalty in performance—decreased height of orbit in the case of a satellite—which results from even ounces of weight. A pound of weight in the satellite is equivalent in lost performance to several hundred pounds of weight in the earlier stages of the launching vehicle.

Power requirements must be reduced to the absolute minimum because of the effects of overheating and the extra weight of the power supply, whether it be chemical batteries or solar cells.

The environment encountered in space flight introduces other problems. Heat must be controlled by a careful balance between radiation and absorption; there is no atmosphere available for convection cooling nor is there a heat-sink available. Erosion of surfaces, particularly those of solar cells, must be checked. If little or no protective coating is required, a considerable weight saving can be effected. The effect of high energy radiation upon components cannot be ignored.

Types of Telemetry Information

Information to be transmitted to the ground falls into many categories. Temperatures at certain critical positions inside the payload, as well as the temperature outside the payload, are needed for future design information. Radiation of many types must be identified and measured in order to determine its effect upon future operations. Micrometeorite particles must be studied to determine size, density, and hazard potential. In fact, all of the environmental conditions are different from anything previously encountered.

As space exploration continues there will be a requirement to transmit pictures of the surfaces of the moon, and of Venus, Mars, and other planets back to earth. The astronomers will observe the stars from an orbiting observatory outside the earth's atmosphere, marking the first time that man has studied the radiation of the stars without the earth's atmosphere as a filter. Pictures of the earth taken from several hundred miles altitude will give meteorological information regarding cloud formations to increase the accuracy of long-range weather forecasting.

When the first manned space flights are made, physiological data must be under almost constant surveillance by medical personnel.

System Requirements

The cost of operations and the nature of the experiments place an imposing list of requirements on the telemetry system. Accuracy of the over-all system, which could be rather crude at first, must be improved to permit accurate measurements of brightness levels of stars, and other high resolution measurements in the order of 0.1 per cent. Bandwidths of 3 to 6 mc will be needed if high resolution TV pictures are to be transmitted in real time. The range of operation for the radio link will vary from a few hundred miles for a low-altitude, short-lived satellite to many millions of miles for deep space shots. Length of operating life will vary widely for different operations. For instance, low-altitude satellites may last only a few hours, while high satellites or space probes may continue to function for years.

The telemetry system used on many payloads should be extremely flexible both in the type of data handled and in data handling rate. A variable data transmission rate will allow more efficient use of radiated power where channel information rates and distance vary over wide limits. For example, with highly eccentric satellite orbits, velocity and necessary information rates are at a minimum when the satellite is at its maximum distance from the earth.

The type of data will vary in some experiments when a single transmitter is used for more than one purpose such as for alternately sending environmental data and a television picture. In addition, the system should be able to switch on command from a low radiated power mode to a high power mode as a deep space probe travels farther away from the earth, or as the bandwidth of the information is increased.

Space Telemetry Systems Previously Used

Table I lists some of the nonmilitary space probes and satellites which were fired previous to the writing of this paper. It provides a ready comparison of pertinent telemeter data, including frequency, modulation scheme, power, and weight, as well as orbital or range data. The subcarrier bands listed are the standard Inter-Range Instrumentation Group (IRIG) bands.

Telemeter transmitters of 10 to 60 mw power output have provided quite adequate signals for data reception to ranges of 2000 to 3000 miles with bandwidths of 2 to 3 kc. Most payloads have used more than one transmitter in order to provide a certain amount of redundancy and to provide additional data channels.

While it is apparent that several different modulation schemes have been used, it should be pointed out that, with the exception of the Vanguard series, telemetry systems have adhered closely to IRIG standards.

Many of the payloads were designed and fabricated on a short time schedule and as a result the organizations supplying them used the telemetry systems with which they were most familiar and which at the same time appeared to be adequate for the scientific experiment. These systems include FM/FM, FM/AM, FM/PM, PCM/PM, and the Vanguard PDM-FM/AM.

TABLE I

Name	Frequency (mc)	Transmitter Power	Type Modulation	Amount of Carrier Modulation	Antenna Type	Radiation Polarization	Subcarrier Bands‡ (If IRIG FM/FM)	Approximate Transmitter Life	Type Power Supply	Weight of Batteries (pounds)	Apogee (miles)	Perigee (miles)	Launch Vehicle
Explorer I	108.00	10–20 mw	FM/PM	0.7 radian (rms)	Dipole	Linear	2, 3, 4, 5	3½ months	Mercury batteries	2.2	1,573	224	Jupiter C
	108.03	10 mw	FM/AM	50 per cent	Turnstile	Circular	2, 3, 4, 5	½ month		2.1			
Explorer III	108.00	10 mw	FM/PM	0.7 radian (rms)	Dipole	Linear	2, 3, 4, 5	2½ months	Mercury batteries	2.2	1,746	121	Jupiter C
	108.03	*60 mw	FM/AM	50 per cent	Dipole	Linear		1½ months		3.1			
Explorer IV	108.00	10 mw	FM/PM	0.3 radian (rms)	Dipole	Linear	1, 2, 3, 4, 5	1¾ months	Mercury batteries		1,380	163	Jupiter C
	108.03	30 mw	FM/AM	100 per cent	Dipole	Linear	1, 2, 3, 4, 5	2¼ months					
Explorer VI (Paddlewheel)	108.06	*60 mw	FM/FM		Two monopoles	Linear	1, 2, 3, 4, 5, 8	1 month	Solar cells and chemical batteries	17	26,000	156	Thor-Able
	108.09	*60 mw	FM/FM		Two monopoles	Linear	1, 2, 3, 4, 5, 6						
Vanguard I	378.00	* 5 watts	PCM/PM		Two monopoles	Linear							
	108.00	10 mw	FM	Approx. 6 kc	Turnstile	Circular		19 days	Mercury batteries	10.5 ounces	2,453	407	Vanguard
Vanguard II	108.03	5 mw	FM	Approx. 6 kc	Dipole	Linear		Years	Solar cells				
	108.00	10 mw	FM	Approx. 6 kc	Turnstile	Circular		27 days	Mercury batteries	10.5 ounces	2,061	350	Vanguard
Vanguard III	108.03	* 1 watt	AM/AM	60 per cent				18 days		8½			
	108.00	30 mw	PDM-FM/AM	100 per cent	Turnstile	Circular		90 days Predicted	Silver—zinc	22.35†	2,330	318	Vanguard
Pioneer I	108.03	*80 mw	AM	100 per cent	Turnstile	Circular							
	108.06	300 mw	FM/PM	1.0 radian	Dipole	Linear	1, 2, 3, 4, 5, 6	43 hours	Mercury batteries		Altitude 70,700	Space probe. Did not orbit	Thor-Able
Pioneer II	108.09	100 mw	FM/PM	1.0 radian	Dipole	Linear	1, 2, 3, 4, 5, 6	Life of probe					
	108.06	300 mw	FM/PM	1.0 radian	Dipole	Linear	1, 2, 3, 4, 5, 6	Life of probe	Mercury batteries		Altitude 963	Space probe. Did not orbit	Thor-Able
Pioneer III	108.09	100 mw	FM/PM	1.0 radian	Dipole	Linear	1, 2, 3, 4, 5, 6						
	960.05	180 mw	FM/PM	0.79 radian (peak)	Conical payload served as the antenna	Linear	1, 2, 3	38 hours	Mercury batteries	6.6	Altitude 63,580	Space probe. Did not orbit	Juno II
Pioneer IV	960.05	180 mw	FM/PM	0.79 radian (peak)	Conical payload served as the antenna	Linear	1, 2, 3	90 hours	Mercury batteries	6.6	In orbit around the sun. TM data and tracking were continuous to 400,000 miles.		Juno II

* Transmitters have command features.
† This battery weight includes power for the magnetometer polarizing field.
‡ IRIG FM/FM Subcarrier Bands (cps) (in part): 1) 400, 2) 560, 3) 730, 4) 960, 5) 1,300, 6) 1,700, 7) 2,300, 8) 3,000.

While some satellites have used continuous transmission of data, it has been found more practical in other instances to store data for subsequent readout by radio command. By this approach it is possible to economize both on transmitter power and on the number of ground receiving stations. A tremendous number of stations would be required to give full-time global coverage of satellite operations.

Data storage has taken two general forms. On Explorer III and Vanguard II, data for a complete orbit were recorded on tape and played back upon command during passage over a Minitrack station. Vanguard III used magnetic-core storage devices in which the flux level was a function of the peak value of current flow in the input winding. Two such devices were switched alternately between storage and readout conditions once each orbit when the satellite passed from shadow into sunlight, thus providing a reading of the maximum input encountered on the previous orbit.

VANGUARD SYSTEM

The PDM-FM/AM system which was designed for use on the later Vanguard satellites during the International Geophysical Year will also be used on some of the satellites to be placed in orbit with the Scout vehicle. Forty-eight information channels are nominally provided consisting of 16 FM tone bursts and 32 PDM time-interval channels. Information is carried by three methods:

1) frequency of each tone burst;
2) duration of each burst;
3) time interval between bursts.

Both the data sampling rate and the number of data channels are readily changed. Data sampling rates may be changed by varying the maximum time duration of the samples; two or more channels may be paralleled (supercommutation) to increase sampling rates on particular inputs as necessary. The system is so designed that the number of data channels may be readily changed in multiples of three by the removal or addition of circuit sections as needed. The system's 108-mc transmitter is amplitude modulated.

A typical Vanguard telemetry system, including transmitter, encoder, and batteries, weighed approximately six and one-half pounds and occupied a volume of 510 cubic inches. The basic 48-channel encoder weighed approximately 3.2 ounces and consumed 12 mw of battery power. The 108-mc transmitter was designed to radiate 80 mw of power. The low power requirements permit continuous operation for a period of three to four weeks on the mercury batteries contained in the package size and weight given above. In the typical Vanguard system the tone bursts were in the frequency range of 5 to 15 kc, with the time durations of the bursts and times between bursts falling between 5 and 30 msec. Thus, all 48 channels were sampled three to four times per second depending upon the values of the input signals to the PDM channels. The encoder was designed to accommodate variable voltages (0 to 5 volts, variable currents (typically one milliampere full scale) or variable resistances (5000 ohms maximum).

FM SYSTEM

The FM/FM, FM/PM, and FM/AM systems used on the Explorer and Pioneer satellites and space probes have used standard IRIG, $7\frac{1}{2}$ per cent frequency modulated subcarrier channels. The outputs of the subcarrier oscillators have been combined as in a standard FM/FM telemetry system and this combined output has been used to modulate the transmitter carrier frequency. Pioneers have all used phase modulation of the carrier, while Explorer payloads have used either frequency, phase, or amplitude carrier modulation. The receiving equipment of the original ground stations was designed with narrow bandwidths in order to provide minimum noise for maximum range of telemetry coverage. This narrow bandwidth capability has required that the subcarrier oscillators be confined to the lower frequency bands, as shown in Table I.

The telemetry system used on the Explorer I was typical of those used on the Explorer I, III, and IV satellites. A volume of 120 cubic inches contained the complete telemeter unit. Weight of the electronic payload was 11.8 pounds, and total equipment weight was 18.2 pounds, including the structure and antenna. Four subcarrier oscillators operating in bands 2, 3, 4, and 5 were used. The data, which included skin and internal temperatures, cosmic ray measurements, and micrometeorite impact data, were duplicated on two transmitters. One at 108.00 mc was a 10- to 20-mw transmitter employing phase modulation and the second was a 60-mw amplitude modulated transmitter. The 10-mw transmitter was powered by six RM-42 mercury cells and the 60-mw transmitter by 24 RM-42 mercury cells.

In the Pioneer I and II space probes the electronic payload weighed 22.5 pounds, including two telemetry transmitters, command receivers, antenna, batteries, and other electronic equipment. One transmitter radiated 100 mw; the other, 300 mw. On command from the ground station the 300-mw transmitter functioned as the return link of a two-way Doppler system. Both transmitters were phase modulated with information from 6 subcarrier oscillators in bands 1 through 6. Power was obtained from mercury batteries which were expected to provide a life of 100 hours for the low-power transmitter and of 10 to 15 days for the high-power transmitter.

The Pioneer III and IV space probes used a single FM/FM transmitter operating at 960.05 mc with a radiated power of 180 mw. Three subcarrier oscillators in bands 1, 2, and 3 were used. Power was obtained from 6.6 pounds of mercury batteries. The payload of Pioneer IV was tracked and telemetry data received until battery exhaustion occurred 82 hours after launching, at a distance of 407,000 miles from the earth

PCM SYSTEM

The PCM/PM system is used on the Explorer VI (Paddlewheel) satellite. This system transmits frames of information which repeat periodically. Each frame consists of eleven words, with one word being used for synchronization and ten words for data. Each word is made up of twelve binary bits, two bits for word synchronization, and ten bits for data. In order to conserve power when rapid data transmission rates are not required, the system is designed to operate at rates of 1, 8, or 64 bits per second. A command signal from the ground turns on the telemeter link and determines the bit rate. The RF carrier is phase modulated.

SOVIET TELEMETRY

The satellites put into orbit by the Soviet Union have used transmitter frequencies of 20.05 mc and 40.01 mc. These low frequencies have been useful in certain radio propagation studies and have provided signals in frequency bands covered by the radio receivers used by radio "hams" throughout the world. This latter fact was a great factor in increasing and stimulating worldwide interest in the first launching. In addition to telemetry transmitters in the 20-mc band, the two Soviet lunar probes used a transmitter operating at 183.6 mc.

The Soviet satellites have used both chemical batteries and a combination of chemical batteries and solar cells to power their telemetry systems. Due to the relatively heavy payloads available to them, they have used higher transmitter powers than most U. S. satellites. This power coupled with the lower frequencies has given telemeter coverages over quite large areas from each receiving station.

In contrast to the Soviet systems, U. S. satellites have operated on a higher frequency to take advantage of the increased accuracy of radio tracking obtainable at the shorter wavelengths, and have concentrated on obtaining the maximum amount of scientific information possible within the lesser weight limits available for payloads.

DATA COLLECTION

Telemetry data collection for the nonmilitary program has been accomplished primarily by facilities owned or contracted for by the National Aeronautics and Space Administration, such as the Minitrack network and university groups in the United States and abroad; facilities under the cognizance of Department of Defense groups; the 250-foot-diameter radio telescope at Jodrell Bank, England; and the facilities of various volunteer groups.

MINITRACK NETWORK

The Minitrack network is composed of ten stations which were set up under Project Vanguard to track and record data for satellites fired in low inclination orbits during the IGY. Now a facility of the NASA, this network uses radio interferometry techniques to determine satellite angular position vs time data which are accurate to within 20–200 seconds of arc and 1 msec of time. Tracking capability now exists at 108 mc for U. S. satellites with some secondary capability at 40 mc for use with Soviet satellites. Facilities are presently being installed for use on the new 136–137-mc band allocated for space communications.

Present Minitrack tracking and telemetry stations are located at Blossom Point, Md.; San Diego, Calif.; Fort Myers, Fla.; Antigua, West Indies; Quito, Ecuador; Lima, Peru; Antofagasta, Chile; Santiago, Chile; Woomera, Australia; and Esselen Park, South Africa. Four new stations being added to provide increased coverage for polar orbiting satellites will be located in Minnesota, Alaska, Newfoundland, and Europe.

The telemetry receivers in operation at each of the Minitrack stations are of an amplitude modulation superheterodyne type and were originally intended to receive the Vanguard telemetry signals during the IGY program. Adapters have been added to the receivers for the 108-mc band to permit the reception of frequency modulated and phase modulated signals. These receivers have three bandwidths, selected as desired, of 31, 5.8, and 3.4 kc. Sensitivities at these three bandwidths are approximately −120, −127, and −130 dbm, respectively. Antennas with a gain of approximately 12 db are used for telemetry reception. In addition to the 108-mc band, five stations are equipped to receive and record amplitude modulated telemetry data on 20 and 40 mc. Esselen Park is equipped to handle telemetry at 20 mc only and has no 108-mc capability.

All stations are equipped with command transmitters which are used in conjunction with receivers in certain satellites to turn on data storage devices and telemetry transmitters.

FUTURE SPACE TELEMETRY

Even at this early stage in the exploration of space, the need for many advancements in the fields of telemetry and communications is recognized. Programs such as the meteorological satellite will require transmission of high resolution pictures of the earth's surface. Lunar and planetary exploration programs will require a similar capability over much greater ranges. These and other programs such as active repeater communications satellites will require wide bandwidth data transmission with reliable operation over long periods of time. Astronomical and other sophisticated scientific studies bring additional requirements. Briefly stated, we must develop a capability for reliable transmission of data at much greater rates, over tremendously increased distances and, in many cases, for longer periods of time.

Several possible approaches to solving these problems exist, all of which must be thoroughly explored and exploited.

The "brute force" approach to the problem is to increase the effective transmitter power in order to allow wide bandwidth operation with reasonable signal-to-noise quality. Transmitter power alone is not a practical solution, however. If 5 watts of radiated power are necessary to transmit certain data to the earth from the moon, about 50,000 watts will be required to produce the same results from Venus at its closest approach to the earth. Thus, although higher powered, more efficient transmitters will be developed, further increases in effective transmitted power must come from the use of high-gain transmitting antennas on the vehicle. The development of higher-gain, more efficient ground-based receiving antennas is also desirable.

New and improved power sources will be required because of increased power drain and longer periods of operation. In addition to these factors, solar cells on interplanetary missions will drop to about 45 per cent of normal output at Mars and to about 4 per cent at Jupiter.

In addition, reductions in system noise and improvements in system efficiency are also necessary. Range of operation will be increased by the use of low-noise amplifiers in the ground receiving equipment. Parametric and maser amplifiers are now in their infancy and show great promise along these lines. New improvements of these amplifiers can be expected to be in field use within a relatively short period.

It will be possible to conserve bandwidth by transmitting information on a reduced time scale when information is not required in "real time." However, in all cases a careful application of information theory will be required to determine the best telemetry systems to establish maximum information capacity. Perhaps the greatest step forward in the field of space telemetry will be brought about by careful application of modern information theory to the development of more efficient communications systems.

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Signal-to-Noise Considerations for a Space Telemetry System*

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Summary—**A signal-to-noise comparison is given between the pulse-frequency channels and the pulse-width channels for the FM/PDM-AM telemetry system. It is shown that the pulse-frequency channels have either a higher signal-to-noise ratio or greater information rate capability than the pulse-width channels.**

INTRODUCTION

A CONSIDERABLE effort is being expended towards the development of rocket vehicles capable of propelling large payloads into space. As the payload grows larger, the need for miniaturization decreases and is supplemented by redundancy to improve reliability; however, for some time to come there will be a need for light-weight, low power-drain telemetry systems. A current example of this need is illustrated in the Scout rocket program.

The Scout vehicle is a relatively inexpensive four-stage rocket capable of orbiting payloads in the 150-pound category. This vehicle will be used on a cooperative basis by scientists of other nations. Because of the restricted weight limitations of the payload, it is imperative that the telemetry electronics compose only a small portion of the instrumentation. There are several types of existing systems which meet this requirement, and most of these have enjoyed success in orbiting vehicles.

In general, for those systems in which multiple-channel simultaneous data is needed, at least two different systems are in use. The early Explorer and Pioneer telemetry employed FM subcarrier oscillators which phase-modulated the transmitter carrier. The subcarrier frequencies were derived from the Inter-Range Instrumentation Group Standards and occupied the lower bands. Vanguard III used a different type of system which was designed primarily to fit a particular set of scientific experiments. It is the purpose of this paper to explore the signal-to-noise considerations for this latter type of telemetry, since little information on the subject has appeared in print.[1,2]

PRINCIPLE OF OPERATION

Vanguard III is the combination of two earlier scientific satellites of the Vanguard series: the proton-precessional magnetometer and the X-ray-environmental satellite. The magnetometer has a separate transmitter, and a four-second turn-on is accomplished by command transmitters at each ground station. The X-ray and environmental portion is transmitted continuously except for the four-second interval during command. The circuitry for the continuous telemetry has been described in the literature; therefore, only a discussion of its operation will be given here. The telemetry encoder shown in Fig. 1 has a 48-channel capacity and weighs 3.2 ounces. The power drain is 10 milliwatts.

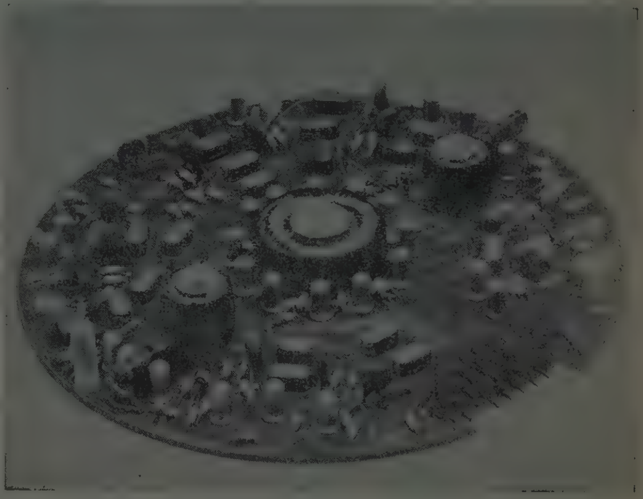

Fig. 1—View of telemetry encoder.

The basic system employs sixteen subcarrier oscillators which are sampled sequentially in time. Each oscillator is deviated over a frequency range of from 5 to 15 kc by its input signal voltage. The composite signal as fed to the transmitter would appear as a train of audio pulses, the frequency of each being representative of a signal level in a particular channel. It is also possible to modulate the pulse widths of each of the audio pulses in proportion to other signal inputs; likewise, the time lengths between adjacent pulses may be modulated in proportion to still other signal inputs, which would extend the number of channels available. As in other systems, the principle of super- and subcommutation may be employed either to increase or decrease the number of channels. Supercommutation is employed in the illustration in Fig. 2 which shows a typical frame of the telemetered signal from a satellite such as Vanguard III. The maximum length of a pulse or the separation between pulses is twenty milliseconds, while the minimum length is four milliseconds. One of the pulse lengths is maintained fixed and longer than any

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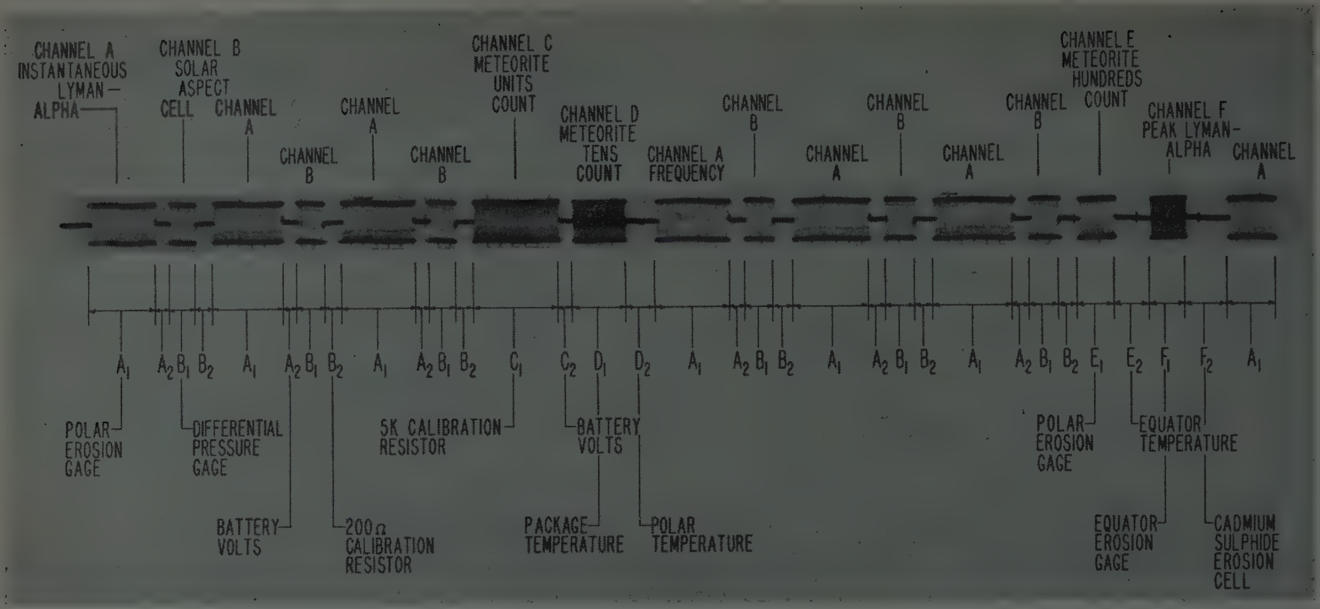
CHANNEL A INSTANTANEOUS LYMAN—ALPHA
CHANNEL B SOLAR ASPECT CELL
CHANNEL A
CHANNEL B
CHANNEL A
CHANNEL B
CHANNEL C METEORITE UNITS COUNT
CHANNEL D METEORITE TENS COUNT
CHANNEL A FREQUENCY
CHANNEL B
CHANNEL A
CHANNEL B
CHANNEL A
CHANNEL B
CHANNEL E METEORITE HUNDREDS COUNT
CHANNEL F PEAK LYMAN-ALPHA
CHANNEL A
A_1 A_2 B_1 B_2 A_1 A_2 B_1 B_2 A_1 A_2 B_1 B_2 C_1 C_2 D_1 D_2 A_1 A_2 B_1 B_2 A_1 A_2 B_1 B_2 E_1 E_2 F_1 F_2 A_1
POLAR EROSION GAGE
DIFFERENTIAL PRESSURE GAGE
BATTERY VOLTS
200Ω CALIBRATION RESISTOR
5K CALIBRATION RESISTOR
BATTERY VOLTS
PACKAGE TEMPERATURE
POLAR TEMPERATURE
POLAR EROSION GAGE
EQUATOR TEMPERATURE
EQUATOR EROSION GAGE
CADMIUM SULPHIDE EROSION CELL

Fig. 2—Typical telemetered signal showing channel allocations.

of the rest, so that synchronization upon decoding may be effected.

The information content of this type of signal is based on the frequency of a pulse being proportional to a signal input and also the duration of a pulse and the time between pulses being proportional to signal inputs. This characterizes two basic types of measurement: frequency and time. Both of these measurements will be described in detail and signal-to-noise error ratios derived for each.

Measurement of Pulse Frequency

The frequency measurement for a single channel may be thought of as a "window" looking in on the signal. The window is "open" only during the times that a particular channel is being sampled. The exact length of time that the "window" is held open is, of course, determined by a signal in the pulse-width channel. If the signal is changing during the time that the "window" is open, the change in pulse frequency is proportional to the slope of the signal at that point while the average frequency denotes the magnitude of the signal. Thus, at each sample point it is possible to get not only the magnitude of the signal but the slope as well. Knowing the slope at each sampling point is equivalent to doubling the sampling rate. The result, then, is that the information bandwidth of the frequency channels is effectively doubled.

To optimize the signal-to-noise ratio, the pulse is passed through a filter. There are several factors which determine the proper width of the filter. First, consider a pulse frequency equal to 10 kc and a width of 10 milliseconds. Suppose that it is desired to detect only the presence or absence of the pulse. The power spectrum of the pulse is given by:

$$\begin{aligned}\Omega(\omega) &= \int_{-\infty}^{+\infty} \epsilon^{-j\omega t} e(t)\,dt \\ &= K\left[\frac{\sin[(\omega-\omega_0)T/2]}{\omega-\omega_0} + \frac{\sin[(\omega+\omega_0)T/2]}{\omega+\omega_0}\right]. \qquad (1)\end{aligned}$$

The major part of the energy is contained in the band plus and minus $1/T$ cps, so that a filter with a bandpass of 200 cps should reproduce the pulse with only a small loss of detail. The rise and fall time (for a simple LC filter) is $1/(2\pi\times 200)$ second or 0.8 millisecond. The bandwidth may be narrowed further. In three time constants the response is up to 95 per cent. With this criterion,

$$3\tau = 10 \text{ milliseconds},$$

or

$$\tau = 3.3 \text{ milliseconds},$$

then

$$f = \frac{1}{2\pi \times 3.3}\text{ kc} = 48 \text{ cps}.$$

There would be, of course, 5 per cent crosstalk into the next pulse channel, which could cause an error. However, if the object is to detect only the presence or absence of a pulse, this could be a satisfactory bandwidth.

The question can now be raised, "How accurate may the frequency of the single pulse be, read in the presence of noise?" There are two ways to approach the problem. As long as the signal amplitude is greater than the noise, a counter may be used to measure the frequency directly, provided that the bandwidth of the filter is wide enough to allow the output to reach steady state. In the example of a 10-kc pulse 10 milliseconds long, a 200-cps filter would be quite adequate to allow the switching transient to damp out. The counter could count for almost the full 10 milliseconds. If the bandwidth were narrowed substantially, there might not be enough counting time in the 10 milliseconds, due to the longer switching transient. It would take 10 milliseconds for the output to build up to essentially full amplitude, at which time the pulse amplitude drops to zero. The filter will continue to ring in a damped oscillation for another 10 milliseconds at the resonant frequency of the filter.

In the presence of noise, the error percentage in the frequency measurement would be:

$$\text{error percentage in frequency} = \frac{N}{S} \times \frac{100}{2\pi f_1 T_1} \qquad (2)$$

where

N = rms noise voltage
S = signal voltage
f_1 = pulse frequency
T_1 = pulse duration.

If the frequency of the pulse is known to have a finite number of discrete values, a bank of narrow filters, each tuned to one of the discrete frequencies, may be used to improve the signal-to-noise ratio. Here the problem reduces to determining only the presence or absence of a signal. The outputs of the filters can be added in logic circuitry in such a way that the greatest signal is the only one present in the combined output.

Measurement of Pulse Width

In this telemetry system the widths of the pulses are modulated; this has a definite effect on the minimum width of filter that can be used. The minimum pulse width is four milliseconds. From the frequency measurement viewpoint the bandwidth should not go below 500 cps because of error introduced by switching transients; however, in reading out the pulse-width information, it is seen that a much wider bandwidth is needed. Consider again a 10-kc pulse 10 milliseconds long. After this is passed through a filter, the leading and trailing edges of the pulse will have a rise and fall time dependent on the filter bandwidth; the narrower the bandwidth, the longer is the rise and fall time. The pulse duration is defined as the time between midpoints of the rise and fall portions. In-phase noise will cause an apparent shift in the midpoint of the leading and trailing edges of the pulse. The amount of the shift is calculated much in the same manner as the effect of quadrature noise on the frequency measurement was calculated. The per cent of error in the width measurement is given by:

$$\text{error percentage in pulse width} = \frac{N}{S} \times \frac{100}{2\pi B_2 T_2} \qquad (3)$$

where

N = rms noise voltage in passband B
S = signal voltage
B_2 = bandwidth of filter
T_2 = maximum change in pulse duration.

Although the expression is almost identical to (2), N contains a factor of $\sqrt{B_2}$ which can be cancelled by the B_2 in the denominator, making the per cent of error inversely proportional to $\sqrt{B_2}$. Increasing B_2 will decrease the per cent of error until the bandwidth becomes so wide that the rise and fall occur in a few cycles of the pulse frequency, f_1. Taking B_2 at this point to equal the pulse frequency, f_1, the ratio of (3) to (2) gives the improvement factor of the frequency measurement over the pulse-width measurement. The improvement factor is found to be

$$\eta = \sqrt{\frac{f_1}{B_1}}$$

where

η = improvement factor
f_1 = pulse frequency
B_1 = bandwidth of frequency measurement.

Conclusion

If the same bandwidth is used for both the width and the frequency measurements, there is little difference in the signal-to-noise ratios. In poor signal-to-noise conditions the bandwidth for the frequency measurement may be narrowed down to get an effective improvement. For this reason signals which may not need the higher frequency response or accuracy are placed on the width channels. The frequency channels have the advantage of giving both the magnitude and the slope of the signal for the same number of sampling points.

The signal is adaptable to either single-sideband or double-sideband suppressed carrier modulation to gain a threshold improvement over amplitude modulation.

The best signal-to-noise ratio is obtained when the oscillators are limited to a finite number of discrete frequency levels and no width channels are used. This type of system would have the same noise threshold as pulse code modulation, with the added advantage that the digits are not restricted to a binary system. The advantage is gained at the expense of distributing the signals over a wider bandwidth, even though the predetection bandwidth is the same in both cases.

Telemetry Bandwidth Compression Using Airborne Spectrum Analyzers*

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Summary—**In practice, most missile test signals requiring wide-band telemetry channels are high-frequency random signals obtained from vibration accelerometers, etc. However, their information content is contained in their slowly-varying statistical parameters. Considered are the unique problems associated with first extracting the spectral density plot of high-frequency random signals (one of the most useful of the statistical parameters), and then presenting this to the telemeter. Design formulas are derived showing the interrelationship between the parameters of the random signal and the characteristics of the telemetry. Formulas indicating the tremendous improvement in bandwidth efficiency are also given.**

INTRODUCTION

A STUDY of telemetry systems used in connection with missile or aircraft development indicates that most of the information being gathered is slowly-varying in nature. Temperatures, pressures, body attitudes, etc., are typical of such quantities. Because these are slowly-varying, they require only a narrow bandwidth in the telemetry link; only a small portion of the data gathered requires wide bandwidth. Most of the wide-band signals that must be telemetered are derived from vibration or acoustic pickups. Now, while their component frequencies may extend to very high limits, the actual information in the signals is contained in their statistical properties, which are varying at a rate comparable to that of the temperatures, pressures, etc., that make up most of the data. It seems, therefore, to be poor telemetry practice to use up almost all of the available bandwidth for these wide-band channels. This bandwidth inefficiency can be eliminated by using small light-weight airborne analyzers, to do the statistical analysis immediately during the flight of the vehicle under study.

This paper discusses the fundamental design considerations of a statistical analyzer to extract in real time the spectral density plot of a random signal (one important type of statistical information) and to present this as a slowly-varying dc voltage to the missile, satellite, or aircraft telemeter. A number of basic problems of some novelty in spectrum analyzer design are introduced by this unique combination of a spectrum analyzer and telemeter. For instance, the conditions of flight are likely to change relatively quickly. Thus, the equipment must be able to do a complete analysis quite quickly. The slow analyzing rate permitted by the continuous recycling of the same data through the analyzer cannot be used. Another problem arises from space, weight, and environmental considerations. The bulky and critically-adjusted components found in most laboratory spectrum analyzers are not acceptable in an airborne telemetry unit subjected to a military or space-flight environment.

It was to explore solutions arising from such unique problems that most of the work outlined in this paper was undertaken.

GENERAL DESCRIPTION OF SPECTRUM ANALYZER

The basic element of the power spectrum analyzer is a narrow band-pass filter, used to divide the frequency bandwidth of the input signal into small intervals. The output of the filter is fed into a detector, which attempts to obtain the average value of some predetermined function of the power, for a specific period of time, in a narrow band of frequencies about the center frequency of the filter. If the bandwidth of this analyzing filter were decreased indefinitely (and the averaging time were also increased indefinitely) the measurement would yield the theoretical power spectral density.

There are two general types of spectrum analyzers: swept types, where one filter is used to sweep over the band of frequencies to be analyzed; and multichannel types where the frequency band is covered by a number, depending on the resolution required, of contiguous filters. The filters are the heaviest and most bulky of the components involved.

The swept type is chosen for study because of its size and weight advantage. Its adaptability to airborne use is enhanced by the fact that it supplies data to the telemeter in a serial fashion. It can be considered to be composed of four principal sections: a swept oscillator-modulator; a narrow-band filter; a squaring circuit; and a detector and integrator.

Let us consider, as an example, a spectrum analyzer which covers the frequency spectrum from 100 cps to 2000 cps. The purpose of the oscillator-modulator is twofold. First, it generates a local signal which is time-modulated by a linear sweep, varying the frequency from 5000 to 7000 cps. Second, it heterodynes this signal with the input signal. A balanced modulator which suppresses the carrier frequency is usually used. As a result, the input signal spectrum is translated along the frequency axis. This frequency translation has the effect of passing the spectrum through the pass band of the analyzing filter at a uniform rate. The oscillator is a symmetrical multivibrator, since the output waveform lends itself ideally to use in a switchtype modulator. In order to insure that the filter analyzes each frequency

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† Ortholog Div., Gulton Industries, Inc., Trenton, N. J.

interval for equal periods of time, the linearity of the sweep is carefully maintained.

A squaring device is necessary to convert the filter output into a signal proportional to power density, if this conversion is required. A linear detector yields root power density. The output signal from the detector is a dc voltage proportional to the desired function of the spectral density. Superimposed on this is a very large random signal, which must be reduced or removed by a low-pass averaging filter if accuracy is desired.

The present study of the analyzer system consists principally of an investigation into its sources of error when the analyzer is operated in real time as part of a telemeter. The problems in mating the analyzer to the conventional telemeter data-gathering system are also covered.

THE ACTION OF THE ANALYZING FILTER

Effect of Sweep Speed

The first problem is that of establishing the general conditions for the satisfactory operation of the analyzing filter, when a rapid analysis is to be carried out. This rapid analysis is one of the principal differences between the present application and the slowly-sweeping laboratory installations. It is important to adhere to major filter properties and to ignore minor refinements. We then stand a good chance of simplifying the filter. Weight and size will thereby be reduced.

An upper limit is set by the bandwidth of the filter, on the speed with which it can be permitted to sweep the spectrum.

It is well known [1] that a tuned circuit of bandwidth B cps, when a sinewave is impressed on it at its center frequency, will build up to the steady-state condition with a time-constant T_b given by

$$T_b = \frac{1}{\pi B}. \tag{1}$$

As the analyzer sweeps across the band, the output of the filter due to a new spectral component must reach steady-state in a small fraction of the time during which the component is in the filter. Otherwise, filter transient conditions will introduce a serious error. The limitation imposed on the rate of frequency sweep, q(cps/sec), can be established by considering the bandwidth B_n (cps) of any "resonant" peak in the spectrum of the random signal.

Obviously, we must keep

$$\frac{B_n}{q} \gg \frac{1}{\pi B} \tag{2}$$

or

$$q \ll \pi B \cdot B_n. \tag{3}$$

If this relationship is maintained, the output of the filter can be considered solely as the steady-state spectral density of the original wave as operated on by the filter pass band. This is important in subsequent work.

Chang [2] considers the effect of the filter sweeping through the spectrum at such a rate that the filter time-constant T_b cannot be completely ignored. He derives results of extreme interest; the filter peak response is depressed by a factor:

$$g = \frac{1}{\sqrt[4]{1 + \left(\frac{4}{\pi}\,\frac{q}{B^2}\right)^2}}. \tag{4}$$

The bandwidth of the filter can be considered as increased by the fraction: $1/g^2$; the center of the filter has shifted by $(4q/B)$ cps. From (4), Chang's criterion for not altering the filter performance is given by

$$q \ll B^2. \tag{5}$$

It will be shown later, that, for accurate measurement, we must have

$$B_n > B.$$

Thus, inequality (5) is a more severe restraint on q than inequality (3). With this restraint, the filter output can be analyzed as a steady-state phenomenon.

Blurring Error

The basic purpose of the spectrum analyzer is to detect the total power (or root power) coming out of the filter. This total power is to be used as a measure of the power spectral density at the filter center frequency. Thus, if there is lack of symmetry in the power contributions obtained from each side of the filter center frequency, an error results. This error is called the blurring error. To hold the blurring error to a minimum, the filter response must be kept symmetrical. But the symmetry of power contributions from the two sides of the filter response can still be upset if the power spectral density changes across the filter band. Chang [2] shows how the blurring error can be estimated. It is given by

$$e_b = \frac{P_w - P_a}{N} \tag{6}$$

where

P_a = the actual power out of the filter,

P_w = the power out of the filter assuming a white-noise source, adjusted to have the same spectral density at $f = f_0$ as P_a, and

N = a factor to normalize the power response of the filter, given by (9) below.

By use of the mean-value theorem, we can transform (6) into

$$e_b = \frac{B^2}{8} \cdot \frac{J}{N} \cdot \left[\frac{d^2 P(f)}{df^2}\right]_{f_0} \tag{7}$$

where

$P(f)$ = the power spectral density under study in watts per cps,

TABLE I

Case	Filter Type	$\lvert Z(u)\rvert$	N	J	J/N	$(B/B_s)^2$
I	Single-tuned	$\frac{1}{\sqrt{1+u^2}}$	π	∞	∞	1
II	Double-tuned (isolated)	$\frac{1}{1+u^2}$	$\pi/2\left(\frac{B_s}{B}\right)$	$\pi/2\left(\frac{B_s}{B}\right)^3$	2.42	0.414
III	Triple-tuned (isolated)	$\frac{1}{(1+u^2)^{3/2}}$	$3\pi/8\left(\frac{B_s}{B}\right)$	$\pi/8\left(\frac{B_s}{B}\right)^3$	1.28	0.26
IV	Rectangular	$\begin{cases}1(-1<u<+1)\\0(u>+1)\\\quad(u<-1)\end{cases}$	$2\left(\frac{B_s}{B}\right)$	$2/3\left(\frac{B_s}{B}\right)^3$	0.33	1

TABLE II

Case	$\left.\frac{e_b}{P_n}\right]_{f_n}$ for $B=B_n/3$	Minimum Acceptance (B_n/B)	
		$e=10\%$	$e=5\%$
I	∞	∞	∞
II	24%	5.0	7.0
III	14%	3.5	4.8

f_0 = the center-frequency of the analyzing filter (cps), and

$$J = \int_{-\infty}^{+\infty} x^2 \cdot Z^2(x) \cdot dx \tag{8}$$

$$N = \int_{-\infty}^{+\infty} Z^2(x) dx. \tag{9}$$

$Z(x)$ is the filter selectivity function (*i.e.*, its amplitude response characteristic).

$$x = \frac{2(f - f_0)}{B}. \tag{10}$$

The ratio (J/N) is calculated for a number of filter types. The results are summarized in Table I; the amplitude characteristics are expressed as $Z(u)$, where

$$u = \frac{2(f - f_0)}{B_s}. \tag{11}$$

B_s is the single-stage bandwidth of the multi-stage filter systems listed.

It is obvious that the blurring error is reduced as the skirt attenuation of the filter becomes more rapid. Only simple filters can be considered as practical for an airborne analyzer. However, a single-tuned filter seems less than acceptable. There is only a minor improvement to be made by using a triple-tuned (isolated) filter over a double-tuned (isolated) one. The rectangular filter is listed only to illustrate the effect of skirt attenuation.

In order to establish the practical impact of the blurring error, let us take (7) and apply it to an isolated spectral peak in the spectral density distribution. Let us assume this peak is obtained by a procedure similar to passing white noise through a single-tuned filter.

Let P_n be the power density at the center frequency f_n of the peak, in watts per cps. Then the spectral density in the neighborhood of f_n can be given by

$$P = \frac{P_n}{1 + y^2} \tag{12}$$

where

$$y = \frac{2(f - f_n)}{B_n}. \tag{13}$$

(B_n is as before the bandwidth of the spectral peak.) Combining (7) and (12) gives, for the blurring error at f_n

$$\left.\frac{e_b}{P_n}\right]_{f_n} = \frac{J}{N}\cdot\left(\frac{B}{B_n}\right)^2. \tag{14}$$

In Table II are calculated the values of (14) for the case when $B = \frac{1}{3}B_n$ for the practical filters of Table I.

Limiting e_b at f_n to a given value establishes the sharpness of the spectral peak that can be handled. Suppose an error limit of $e/100$ is established, then

$$\frac{B_n}{B} \geq 10 \cdot \sqrt{\frac{J}{Ne}}. \tag{15}$$

Calculated in Table II are the limiting ratios established by (15) for the practical cases discussed.

Ten per cent error is quite acceptable for the type of measurement we are making here. The above results indicate that the blurring error only becomes serious when B_n approaches $3B$.

The above calculations compute blurring error as a fraction of the spectral density being measured. However, another effect of the blurring error is the widening of any spectral peaks under study. This widening effect can best be evaluated by using in place of (12) the following formula for the peak:

$$P = P_n \exp\left[-\left(\frac{f-f_n}{m}\right)^2\right]. \tag{16}$$

Hence,

$$\frac{e_b}{P_n} = \frac{-1}{4}\cdot\frac{J}{N}\cdot\left(\frac{B}{m}\right)^2$$
$$\cdot\left[1 - 2\left(\frac{f-f_n}{m}\right)^2\right]\exp\left[-\left(\frac{f-f_n}{m}\right)^2\right]. \tag{17}$$

By putting $P=P_n/2$ in (16), we can obtain B_n:

$$B_n = 2\cdot\sqrt{k_0}\cdot m \qquad (k_0 = \log_e 2). \tag{18}$$

By using, in place of P_n,

$$P_n' = P_n - \left| e_b \right|_{f=f_n} \tag{19}$$

and solving for the value of $(f-f_n)$, making

$$P + e_b = \frac{P_n'}{2}, \tag{20}$$

we obtain the apparent bandwidth B_n' of the spectral peak. The error in bandwidth is therefore

$$\Delta B_n = B_n' - B_n. \tag{21}$$

Let us proceed to do this:

$$P_n' = P_n\left[1 - \frac{1}{4}\cdot\frac{J}{N}\cdot\left(\frac{B}{m}\right)^2\right].$$

To solve for $(f-f_n)$ in (20), we must have:

$$\left(\frac{f-f_n}{m}\right)^2 = \log_e 2 + \log_e\left(1 - \frac{J}{4N}\cdot\frac{B^2}{m^2}\right.$$
$$\left.\cdot\left[1 - \frac{2(f-f_n)^2}{m^2}\right]\right) - \log_e\left(1 - \frac{1}{4}\cdot\frac{J}{N}\cdot\frac{B^2}{m^2}\right). \tag{22}$$

Assuming $B<B_n$ under all conditions, (22) can be approximated as:

$$\left(\frac{f-f_n}{m}\right)^2 = \log_e 2 + \frac{(f-f_n)^2}{2m^2}\cdot\frac{J}{N}\left(\frac{B^2}{m^2}\right). \tag{23}$$

Or,

$$(f-f_n) = \sqrt{k_0}\cdot m\cdot\frac{2}{\sqrt{1 - \frac{J}{2N}\cdot\frac{B}{m^2}}} \tag{24}$$

$$(f-f_n) = B_n\cdot\left[1 + \frac{Jk_0}{2N}\left(\frac{B}{B_n}\right)^2\right]. \tag{25}$$

In this equation we have solved for B_n'. Hence, from (21), we can write that the apparent increase in bandwidth is given by

$$r_b = \frac{\Delta B_n}{B_n} = \frac{Jk_0}{2N}\cdot\left(\frac{B}{B_n}\right)^2 = H\left(\frac{B}{B_n}\right)^2. \tag{26}$$

Table III gives the value of H for the practical filters considered in Table I.

TABLE III

Case	$H(B<B_n)$
I	∞
II	0.83
III	0.44

Eq. (26) evaluates the blurring error as a degradation of the filter resolution. The results tabulated in Table III indicate that we can go to smaller ratios of (B_n/B) for the same percentage error, when we consider the blurring error along the frequency axis, rather than as a power density error. A 10 per cent degradation in resolution does not occur until B_n is decreased to as low as $B_n=2B$.

Once again, the practical blurring error improvement of the more sophisticated circuit of Case III over that of Case II is quite significant.

PROPERTIES OF THE FILTER OUTPUT SIGNAL

It is valid here to assume that the analyzing filter is symmetrical about its center-frequency f_0. It is also valid to assume that its bandwidth satisfies

$$B \ll f_0. \tag{27}$$

Hence the following well-known formula can be used to represent the signal out of the filter:

$$V(t) = C(t)\cos 2\pi f_0 t + D(t)\sin 2\pi f_0 t$$
$$= A(t)\cos\left[2\pi f_0 t + \phi(t)\right] \tag{28}$$

where $C(t)$, $D(A)$, $A(A)$, $\phi(t)$ are slowly-varying functions. $C(t)$ and $D(t)$ are independent.

If the filter is driven by white noise of power spectral density P_1, the autocorrelation function of $V(t)$ is given by

$$R(\tau) = \int_0^\infty P_1 H^2(f)\cdot\cos 2\pi f\tau\cdot df \tag{29}$$

where $M^2(f)$ is the filter power spectral operator.

RIPPLE ERROR USING A SQUARE-LAW DECTECTOR

The filter output signal is applied to a detector, where rectification occurs. The dc level obtained is therefore a measure of the spectral density at f_0. The detector law determines whether power or amplitude, or indeed some other spectral density, is obtained. Due to the beating of the signal components in the detector, a noise signal occurs at the output of the detector, superimposed on the desired dc level. Filtering this new noise spectrum is needed to obtain a useable result. Indeed, the amount of noise passed by the post-detection filter determines the accuracy with which the desired dc output can be read at any instant.

It is desired here to study both outputs of the detector: the dc signal, and the noise spectrum. The post-detector smoothing problem posed by the short sweep time of the analyzer can then be explored. The effect of the design of the frequency-selecting filter on the detector noise spectrum (and hence on the accuracy) will also be established for several common types of filter.

Suppose the signal of (28) and (29) is applied to a detector obeying the law

$$V_0(t) = V^2(t). \tag{30}$$

It can be shown [4] that

$$\overline{V_0(t)} = \overline{C^2(t)} = \overline{V}. \tag{31}$$

Let $P_0(f)$ be the power spectral density of the detector noise. Then it can also be shown [4] that

$$P_0(f) = 4\int_0^\infty (\overline{V})^2 R(\tau)\frac{\cos \pi f \tau}{\cos 2\pi f_0 \tau}\, d\tau. \tag{32}$$

The easiest filter shape to deal with is the rectangular one of Case IV, Table I. This filter, though not practical in this application, will be taken as reference in studying the detector problem. This does not mean that such a filter is the most desirable, as is sometimes implied. It is chosen here as standard primarily because its properties are widely quoted, and it can serve as a base for comparison. With such a filter, we can write:

$$\text{a)}\quad \overline{V}_0 = P_1 B \tag{33}$$

$$\text{b)}\quad R(\tau) = P_1 \frac{\sin \pi B \tau}{\pi B \tau} \cos 2\pi f_0 \tau \tag{34}$$

$$\begin{aligned} \text{c)}\quad P_0(f) &= 0 && f \geq B \\ &= \frac{2(\overline{V}_0)^2}{B}\cdot\left(1 - \frac{f}{B}\right) && 0 \leq f \leq B. \end{aligned} \tag{35}$$

If now the simple single-tuned filter (Case I of Table I), is used, we get

$$\text{a)}\quad \overline{V}_0 = \frac{\pi}{2} P_1 B \tag{36}$$

$$\text{b)}\quad R(\tau) = P_1 e^{-\pi B \tau}\cdot \cos 2\pi f_0 \tau \tag{37}$$

$$\text{c)}\quad P_0(f) = \frac{2}{\pi}\,\frac{(\overline{V}_0)^2}{B}\cdot\frac{1}{\left(1 + \dfrac{f^2}{B^2}\right)}\,. \tag{38}$$

If a synchronously-tuned pair is used (Case II of Table I):

$$\text{a)}\quad \overline{V}_0 = \frac{\pi}{4} P_1 B_s \tag{39}$$

$$\text{b)}\quad R(\tau) = \frac{\pi}{4} P_1 B_s (1 + \pi B_s \tau)\cdot e^{-\pi B_s \tau}\cdot \cos 2\pi f_0 \tau \tag{40}$$

$$\begin{aligned} \text{c)}\quad P_0(f) = \frac{2}{\pi}\cdot\frac{(\overline{V}_0)^2}{B_s}\Bigg[&\frac{\cos a}{(1 + f^2/B_s^2)^{1/2}} \\ &+ \frac{\cos 2a}{(1 + f^2/B_s^2)} + \frac{\cos 3a}{2(1 + f^2/B_s^2)^{3/2}}\Bigg] \end{aligned} \tag{41}$$

where

$$\tan a = f/B_s. \tag{42}$$

An examination of the equations from (33) to (41) reveals two interesting facts:

1) The desired dc output $\overline{V}_0$ is greater by approximately 50 per cent when the simple tuned circuits are used than when the rectangular filter, as defined in Table I, is used.
2) White noise through a rectangular filter detected in a square-law detector has no frequency components above B cps. The other cases produce harmonics up to infinity, although above approximately $f = 2B$ the magnitude of these harmonics is insignificant.

In order to extract the dc voltage from the detector with any accuracy, it is necessary that the cutoff frequency of the smoothing filter be much less than B. In practice, this means that RC filters must be employed, as suitable inductances would prove far too bulky. In this paper, a simple single-stage RC filter is assumed. Then, the rms ripple error, normalized against $\overline{V}_0$, at the output of the filter of time-constant T_c, is given by

$$r = \sqrt{\frac{1}{(\overline{V}_0)^2}\int_0^\infty \frac{P_0(f)df}{1 + 4\pi^2 f^2 T_c^2}}\,. \tag{43}$$

Since we must choose $T_c \gg (1/B)$, we can replace $P_0(f)$ by $P_0(0)$, and bring it outside the integral. Table IV tabulates the value of r for Cases I, II, and IV.

It is obvious from Table IV that smoothing the detected output (square-law) of a rectangular filter has considerably more rms ripple than the other two filters considered, for the same bandwidth and smoothing time-constant. For the practical filters there is only a minor degradation over using the simplest filter (Case I). This is an important result, since we are striving to keep T_c as small as possible (for reasons that will be discussed below), and the resolving filter as simple as possible.

TABLE IV

Case (as classified in Table I)	Smoothing Filter	r	Ratio
IV	Rectangular Filter	$\dfrac{1}{\sqrt{2BT_c}}$	1
I	Single-Tuned	$\dfrac{1}{\sqrt{6.3BT_c}}$	0.55
II	Dual single-tuned (isolated)	$\dfrac{1}{\sqrt{4.00BT_c}}$	0.71

USE OF A LINEAR DETECTOR

A study of (33), (36), and (39) shows that for each type of resolving filter chosen:

$$\overline{V}_0 = (\text{const}) \cdot P_1 \tag{44}$$

where P_1 can be interpreted as the power density at the center of the filter. If a linear detector were used instead of the square-law one discussed, (44) would be replaced by

$$\overline{V}_0' = (\text{const.})\sqrt{P_1}. \tag{45}$$

This is important to the designer of an airborne spectrum analyzer for several reasons:

1) The dynamic range in db required of a telemeter to transmit $\overline{V}_0$, for the same range of P_1, if (45) is used, is only half that required using (44).
2) As the analyzer sweeps across a peak in the spectral density plot, $\overline{V}_0'$ changes at a slower rate than $\overline{V}_0$, and so the frequencies in the signal presented to the telemeter are lower, and more bandwidth conservation is achieved.
3) The derivation of (44) can be done on the ground by squaring the telemeter output. This means that the size and weight of the airborne device is somewhat reduced.
4) The noise ripple error is reduced significantly.

These points will be discussed in more detail below. They are only listed here to indicate the need for exploring the linear detector.

We can deal with the ripple error of the linear detector very quickly [4]. Suppose the signal of (28) and (29) is applied to a detector obeying the law:

$$V_0(t) = A(t). \tag{46}$$

Then it can be shown that

$$\overline{V_0(t)} = \overline{A(t)} = \overline{V}_0'. \tag{47}$$

$P_0'(f)$, the power spectral density of the noise at the output of the linear detector is given by

$$P_0'(f) = \tfrac{1}{4}P_0(f)$$

where $P_0(f)$ is defined by (32), *et. seq.*

Therefore, to convert the ripple errors (r) of Table IV into the errors obtained using a linear detector, with the same resolving filter bandwidth (B) and smoothing time-constant (T_c), we would reduce each value of r by $\frac{1}{2}$. For example, for Case II, the new ripple error would be

$$r' = \frac{1}{2\sqrt{4.00BT_c}}. \tag{48}$$

If the signal received at the ground station is squared, to obtain P_1, the ripple error in the *output* of the squarer would be as given in Table IV once again. This can be seen by letting $E(t)$ be the noise voltage riding on $\overline{V}_0'$ at the ground station. The normalized rms value of $E(t)$ is r'. The input to the squarer, therefore, is

$$\overline{V}_0'(1 + E(t)).$$

The output from the squarer is

$$[\overline{V}_0']^2(1 + 2E(t)).$$

The normalized rms value of $2E(t)$ is $2r'$. This is precisely the same error that would have been obtained using a square-law detector in the airborne spectrum analyzer.

TRANSIENT CONSIDERATIONS—OUTPUT SMOOTHING CIRCUIT

It would seem from the formulas of Table IV, and from (48), that the time-constant of the output smoothing circuit T_c should be as large as possible. There is, however, a limitation on the size of T_c that can be used. This limitation is imposed by the rate with which a spectral peak is being scanned by the analyzer.

As the peak is scanned, the dc average output $\overline{V}_0$ (using a square-law detector) or $\overline{V}_0'$ (using a linear detector) rises and falls with the peak. The smoothing circuit must pass this transient. For the sake of analysis, let us suppose the spectral peak under study is given by (16). As the spectrum is swept, the transient applied to the output smoothing circuit will have a Gaussian shape. For a square-law output, the transient will have the form

$$V = V_m e^{-(qt - f_n/m)^2}. \tag{49}$$

Our objective is to determine the increase in the width of this transient at its 3-db points as it passes through the smoothing circuit. This can be done by following the method developed by Huggins [6]. We note the similarity of form of the convolution integral, when a transient of (49) is applied to an RC smoothing circuit, to that of the probability integral of the sum of two independent variables. Therefore, we represent 1) the input transient (49), and 2) the impulse response of the RC network, as two independent probability distributions of time. The variance about the mean of the impulse response is T_c^2. Therefore, the variance about the mean of the output of the smoothing circuit is given by

$$m_0^2 = m^2 + T_c^2 q^2. \tag{50}$$

Assuming

$$T_c < m, \tag{51}$$

$$m_0 = m\left[1 + \frac{T_c^2 q^2}{2m^2}\right]. \tag{52}$$

Using (18), we can write the normalized increase in bandwidth as

$$r_d = \frac{m_0 - m}{m} = 2k_0\left(\frac{T_c}{T}\right)^2\left(\frac{f_b}{B_m}\right)^2. \tag{53}$$

For a linear detector, (49) is replaced by

$$V' = V_m' \, e^{-(qt - f_n/\sqrt{2}m)^2} \tag{54}$$

and r_d is replaced by

$$r_d' = k_0\left(\frac{T_c}{T}\right)^2 \left(\frac{f_b}{B_m}\right)^2. \tag{55}$$

TELEMETRY CONSIDERATIONS

An important part of the design of an airborne spectrum analyzer system is to make sure that the output of the analyzer is compatible with the telemetry scheme. Let us consider a spectral peak as represented by (49). To reproduce this spectral peak at the ground station of a time-multiplex system, we need at least three samples between the $V_m/2$ points; *i.e.*, if s is the sampling rate (per second), we must have

$$s\ (\text{Sq. Law Detector}) \geq 6\sqrt{k_0}\left(\frac{q}{B_m}\right). \tag{56}$$

Similarly, if a linear detector is used, the spectral peak is represented by (54), and

$$s\ (\text{Lin. Detector}) \geq \sqrt{\frac{k_0}{2}} \cdot \frac{6q}{B_m}\,. \tag{57}$$

Eqs. (56) and (57) can be reinterpreted in terms of bandwidth as follows (f_t = the telemetry bandwidth required to pass the output of the analyzer):

$$f_t\ (\text{Sq. Law Detector}) = \frac{3\sqrt{k_0}}{B_m} \cdot q \tag{58}$$

$$f_t\ (\text{Lin. Detector}) = 3\sqrt{\frac{k_0}{2}} \cdot \frac{q}{B_m}\,. \tag{59}$$

The dynamic range limitations imposed on the analyzer by the telemetry are also worth a brief look. Suppose that the accuracy of the telemetry is such that

$$D = 10 \cdot \log_{10}\left(\frac{\text{maximum signal before overloading}}{\text{minimum useful signal}}\right). \tag{60}$$

The value of D establishes the useful dynamic range of the spectrum analyzer. Unfortunately, the effect of D varies from spectrum to spectrum, and is difficult to evaluate without establishing a specific spectral density plot. However, it is possible to compare the effects of linear and square-law detectors for the same spectrum. Let us assume a white noise signal is being analyzed, and let d equal the variation in power (in db) over which the combined spectrum-analyzer telemeter system gives useful results at the ground station.

Then for the square-law detector,

$$d = D. \tag{61}$$

Let d' = the tolerable variation in power (db) for the linear detector. Then

$$d' = 2d. \tag{62}$$

Eq. (62) is another reason for favoring a linear detector in the airborne unit.

CHOICE OF PARAMETERS

From the above, we can work out a compatible set of basic parameters for any spectrum analyzer telemetry system. There are four main considerations which affect the choice of spectrum analyzer and telemeter parameters:

1) the compatibility, frequency-wise, of the telemeter and the analyzer output [(56) to (59) apply];
2) the choice of T_c small enough to keep the bandwidth error in the smoothing circuit within satisfactory limits [(53) and (55) apply];
3) the choice of B small enough to keep the filter blurring error within satisfactory limits [(26) applies, if we are interested in interpreting the blurring error as a degradation of bandwidth; (14), if as an error in power density]; and
4) the choice of the product $(B \cdot T_c)$ large enough to keep the ripple on the output to a satisfactory level [the formulas of Table IV apply, for a square-law detector; these results can be modified for a linear detector, as illustrated by (48)].

From the above four considerations, the following parameters are chosen to be compatible: a) the telemetry bandwidth f_t or sampling rates, b) the output RC smoothing time-constant T_c of the analyzer, c) the bandwidth of the analyzer B, d) the total frequency band covered by the analyzer f_b, and e) the time for one sweep period T. Other considerations, affecting gain and dynamic range, have also been covered above. However, these are relatively straightforward, compared to 1), 2), 3), and 4) above.

EXAMPLE

An example is perhaps the best way to show how one can try to establish a consistent set of parameters in a systematic manner, for a case II filter.

Let us consider the case where the analyzer has a linear detector only, and where the telemeter is a time-multiplex system. Let us also assume we are interested in the blurring error as a degradation of bandwidth. Let T be the sweep period (in seconds). Then

$$q = \frac{f_b}{T}\,. \tag{63}$$

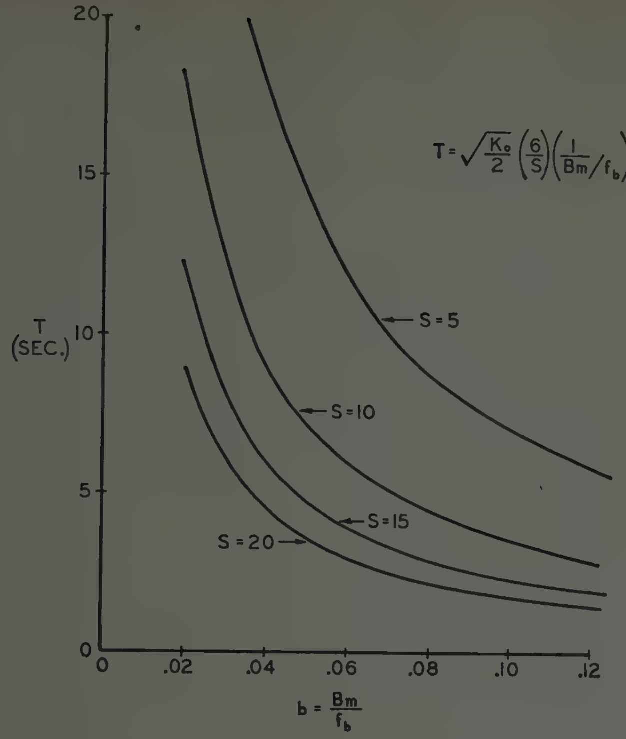

Fig. 1.

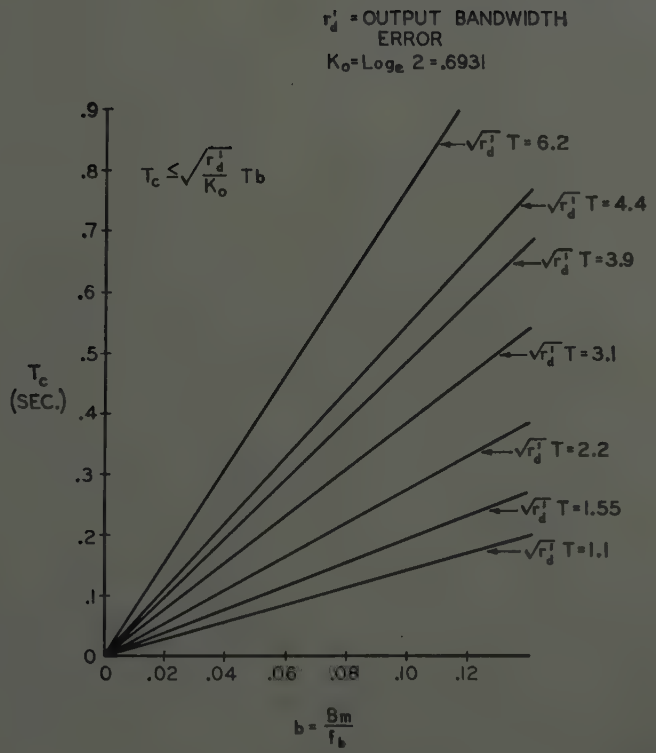

Fig. 2.

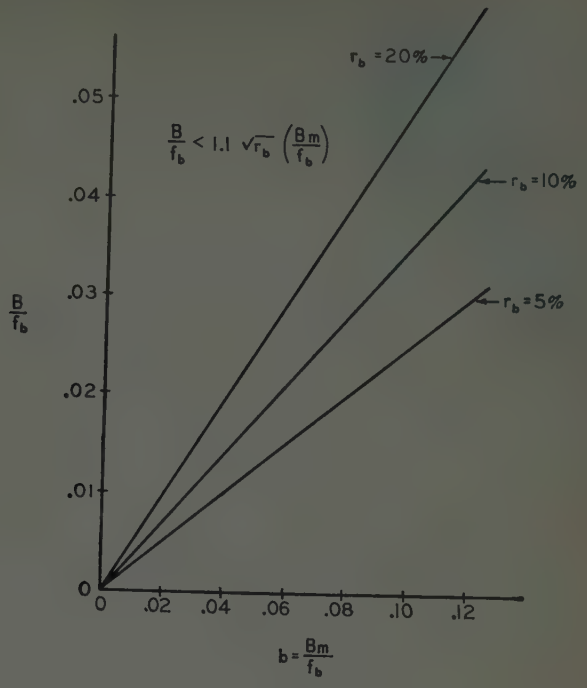

Fig. 3.

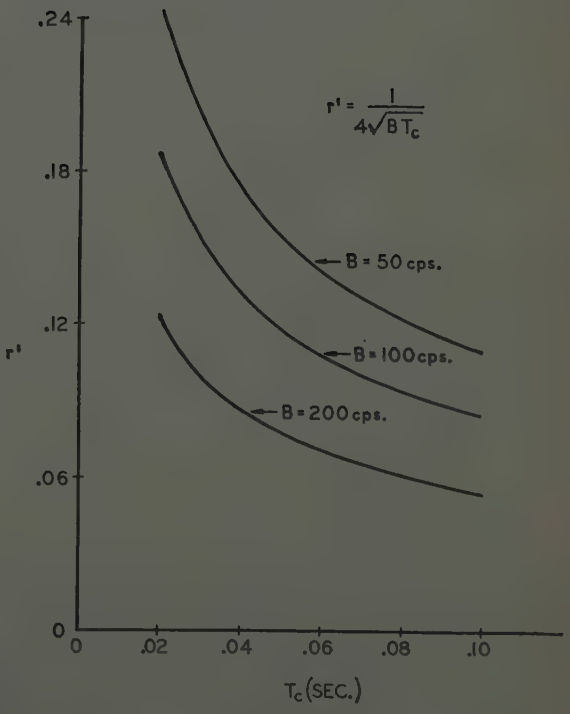

Fig. 4.

and it is possible thereby to eliminate q from all pertinent equations.

It is necessary to establish initially: 1) the maximum ripple error that is tolerable; 2) the maximum output bandwidth error that is tolerable; 3) the maximum blurring error that is tolerable; 4) the sample rate of the time-multiplex system (this is usually well-known; however, the sampling rate can be increased by cross-strapping the multiplexing switch); and 5) the bandwidth of the narrowest spectral peak expected in the random signal under study.

The last, when expressed as a fraction of the total frequency range, is an important quantity. Consequently, let us define it by:

$$b = \frac{B_m}{f_b}. \tag{64}$$

We use the equation for the sample rate, (57), to determine the period of one sweep. This equation is rewritten

$$T \geq \sqrt{\frac{k_0}{2} \cdot \frac{6}{sb}}, \tag{65}$$

and is plotted as Fig. 1.

We can now determine T_c using (55) for the output bandwidth error. This equation is rewritten as

$$T_c \leq \sqrt{k_0^{-1} r_d'} \cdot T \cdot b. \tag{66}$$

Eq. (66) is plotted in Fig. 2.

Next, we use (26) to establish the filter bandwidth B, as follows:

$$\frac{B}{f_b} \leq 1 \cdot 1 \cdot \sqrt{r_b} \cdot b \tag{67}$$

where

$$r_b = \frac{\Delta B_m}{B_m}. \tag{68}$$

Eq. (67) is plotted in Fig. 3.

Finally, we insert the result of (66) and (67) into (48), plotted as Fig. 4. The ripple error so obtained must be less than that desired for a compatible design. If it is not, we must modify our original set of assumptions. The curves given indicate clearly how the various parameters interact, and thus assist in achieving a compromise.

Suppose the frequency band of interest is 2 kc, and we wish an accurate spectral root power density plot for all spectral peaks wider than 200 cps. Suppose also that a 10 sample per second time-multiplex telemeter is to be used. Thus $b = 0.10$, $s = 10$. From Fig. 1, we see the sweep period T must be at least 3.6 seconds. (This also implies that the random signal under study must remain stationary for at least 3.6; if it is not stationary for 3.6 seconds, Fig. 1 will have to be used to select a larger value of s. Cross-strapping in the telemeter will permit values of $s = 20$, 30, 40, etc., to be chosen easily.)

With $T = 3.6$ seconds, and $s = 10$, there will be a telemetry sample every 55 cps across the frequency band f_b. It seems reasonable to require that the output bandwidth error be $r_d' \leq 0.1$. From Fig. 2, we see we must choose $T_c \leq 0.12$ second. Let us also permit the blurring error to be 10 per cent. From Fig. 3, $B/f_b \leq 0.059$. Thus, $B \leq 80$ cps.

From Fig. 4, the ripple error r' is 10 per cent. If this value is too great to be tolerated, some change in the initial choice of parameters is indicated. s or T must be increased, or r_b and r_d relaxed, or a combination of these changes carried out. For most practical cases, however, a value for r' of 10 per cent is very satisfactory. 10 per cent to 20 per cent is usually an acceptable range of values.

It may be of more use in some situations to use values of B and/or T_c which are smaller than the values chosen, accepting the increases in errors that result. For instance, if the data under study cannot be expected to remain stationary for even 3.6 seconds, a smaller value of T must be chosen. This forces T_c to be smaller than 0.12 second; the larger errors that result in this case are not peculiar to the present analyzer, but are the penalty one must pay with any type of analyzer when the amount of data is reduced.

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The Telemetry and Communication Problem of Re-Entrant Space Vehicles*

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Summary—**The problems associated with propagation of electromagnetic telemetry and communication (and guidance) signals through the densely ionized region (plasma) generated by, and surrounding a hypersonic space vehicle re-entering the earth's atmosphere are reviewed. The nature and extent of the plasma-radio wave interference is explained in terms of the classical treatments which have evolved from ionospheric studies. The limitations of the simple classical models are pointed out in the light of the more complex nature of the hypersonically generated plasma. Methods for overcoming the deleterious influences of the plasma are presented, culminating in presentation of some new approaches to this problem.**

INTRODUCTION

A CONCENTRATED effort by space vehicle researchers, and telemetry and communications engineers, is presently being devoted to establishment of means which will permit reliable and continuous RF telemetry, communication, and guidance information to be exchanged between a hypersonic re-entry vehicle and one or more ground stations. For the scope of this paper, hypersonic vehicles may generally be defined as those whose re-entry flight regime lies in the Mach 10 to Mach 20 region. Their particular configurations, whether they be missile nose cones (Jupiter, Atlas, Thor) [1], [2], aerodynamic lift vehicles (X-15 [2], Dyna Soar [3], [4]), manned re-entry capsules (Mercury) [5], or futuristic rockets and manned satellites, are not of concern here. The common element of interest is their hypervelocity during re-entrant flight phases within an atmosphere. These hypersonic velocities during atmospheric flight result in tremendous aerodynamic heating of the vehicle and its immediate surroundings. Among the many effects of the heating is thermal ionization of the atmosphere surrounding the vehicle. This action results in liberation of free electrons and positive ions in the form of a plasma or sheath surrounding the re-entry vehicle, which serves to interfere with the passage of electromagnetic waves. Such interference is characterized by attenuation, refraction, phase shift, and at times complete absorption or reflection of a passing signal. It is for these reasons that the re-entry telemetry and communication problem has commanded so much attention.

Loss of telemetry signals is commonly experienced during specific portions of the re-entry flight profiles of missile nose cones. This situation is serious in that much important information is lost and continuous tracking of the vehicle becomes difficult. Since the time duration of re-entry is small for re-entrant nose cones, loss of signal during this time can be overcome by storage and recording. However, a high price is paid in the form of weight and instrumentation penalties, and recovery of the nose cone at the termination of its flight is mandatory.

When one considers the re-entry paths of manned vehicles such as returning rockets and satellites of future generations, or the Dyna Soar, Mercury, and advanced X-15 vehicles now in the design stage, the RF signal propagation problem assumes proportions far more serious than is present in nose cone telemetry. This is due to the fact that 1) the periods of time during which propagation of RF signals will be blocked by the plasma sheath surrounding the manned re-entry vehicle will be appreciably longer than for the missile re-entry case—probably of the order of hours rather than fractions of minutes; and 2) these vehicles, being manned and in serious need of continuous exchange of guidance and control information, cannot rely upon storage or delayed relay of that information.

In view of these considerations, the importance of devising means of reliably and continually propagating (or propagating at will) telemetry, communication and guidance information to and from vehicles traveling at hypersonic velocities within the earth's atmosphere is self-evident.

In order to arrive at an understanding of the manner in which reliable and continuous signals may be propagated under these conditions, it is first necessary to become familiar with the nature of the environment induced by hypersonic vehicles, and the mechanisms of interaction between the ions of the plasma and a passing electromagnetic wave. From a consideration of these, it then becomes possible to devise means of removing the interfering obstacles or diminishing their effectiveness, and even to arrive at methods of altering the usual behavior of the interfering mechanisms in order to achieve a cooperative influence which can conceivably result in enhancement of the passing RF wave.

NATURE OF THE INDUCED ENVIRONMENT

A re-entrant vehicle moving at hypersonic velocities within the earth's atmosphere produces a complex flow pattern in the surrounding medium. Macroscopically, the resultant shock wave is characterized by a sharp demarcation or front, behind which air pressure and density increase many times over their normal ambient values. These physical changes are attributable to the vehicle velocity within the medium; the resultant frictional forces induce thermal increases which give rise to

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chemical changes in the medium such as dissociation and ionization.

A qualitative presentation of a typical shock front, its associated layers, and their combined relation to the hypersonic vehicle is depicted in Fig. 1. It is emphasized that the illustration is mainly descriptive and is not meant to depict quantitative geometric relationships. The shock front is the strong line of demarcation separating the ambient atmosphere from the induced environment close to the vehicle. In hypersonic flow past a blunt body, the shock front lies close to the surface, tending to wrap itself about the body [6]. The greatest increases in temperature, pressure, and air density (which we shall conveniently refer to here as the "gas parameters") occur behind the portion of the shock in the vicinity of the vehicle leading edge [7]. The extent of these increases is dependent upon vehicle velocity, configuration, and particular characteristics of the natural environment [8]. For re-entrant and high-velocity vehicles such as X-15 or Dyna Soar, gas parameter increases of one or two orders of magnitude over normal values are to be expected [9]. A more detailed presentation of these variations is presented in Fig. 2.

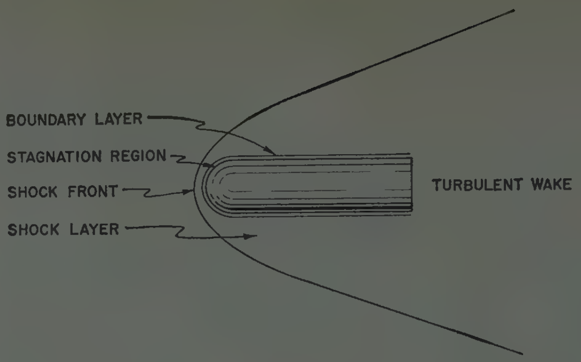


Fig. 1—Geometry of environment induced by a blunt hypersonic body.

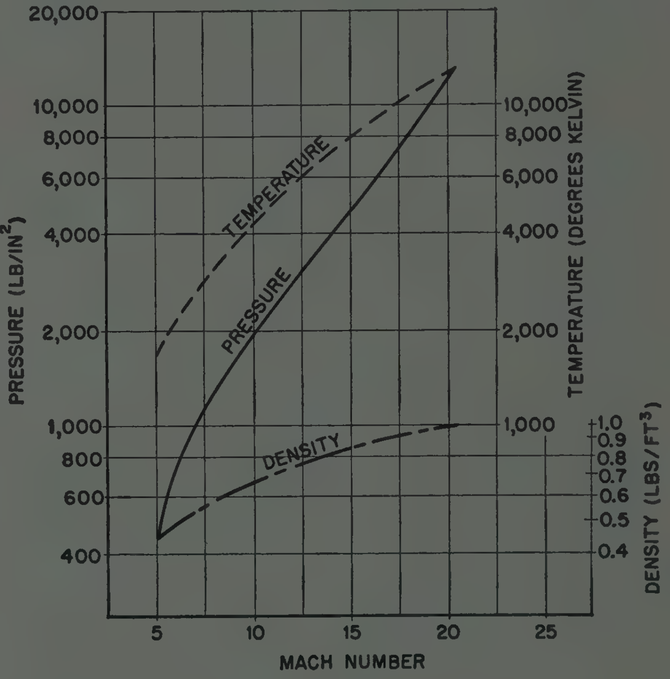


Fig. 2—Induced environment characteristics as a function of Mach number.

The region behind the shock front is termed the shock layer. That portion of the layer immediately adjacent to the vehicle leading edge is referred to as the stagnation region. Here the gas flow is subsonic and temperature, pressure, and density reach their maximum values. Further aft, gas expansion occurs, the flow becomes supersonic, and the gas parameters decrease. The boundary layer is that viscous region through which heat transfer and skin friction occur. On a slender after-body, this layer tends to be thicker at hypersonic speeds than at low speeds. It interacts with the external flow, and this gives rise to an induced pressure with a negative gradient which thins the boundary layer somewhat [10].

The chemical constitution of the atmosphere in the various flow and layer regions of the plasma sheath serves to affect strongly the behavior of electromagnetic signals propagated through them. This comes about because conversion of flow energy into internal energy of the surrounding air results in dissociation and ionization of the gas molecules. Notable among the dissociation products is the formation of nitric oxide which is easily ionized, thus contributing to the formation of large numbers of ions and free electrons [11].

It is this concentration of ionized particles about the hypersonic vehicle which works to interfere with the passage of electromagnetic RF waves to and from the vehicle. This interference is characterized by attenuation, refraction, and possible reflection of the wave during its travel through the region.

The large temperature gradients existing within the boundary layer region bring about diffusion and recombination on the vehicle surface of free atoms from the hot gases surrounding the vehicle. In this process the constituents of the plasma give up their heat of dissociation, thereby transferring heat from the gaseous surroundings to the vehicle surface. This serves to alleviate by a small but insufficient amount some of the interference phenomena mentioned previously.

It is thus seen that the environment induced by a hypersonic vehicle is most complex, and the chemical and physical processes which induce and sustain it in its state of dynamic flux also act to alter the behavior of co-existing mechanisms such as heat transfer, vehicle erosion, electromagnetic radiation, etc. As will be seen later, the complexity of these interrelations and the nature of the combined mechanisms make the accurate expression and analyses of these phenomena a formidable problem.

INTERACTION OF INDUCED ENVIRONMENT WITH ELECTROMAGNETIC RADIATION

The interactions of electrons and waves, and the resultant effects upon electromagnetic propagation have been ascertained by various investigators. Baker and Rice [12], Hartree [13], Lorentz [14] and others have derived relations governing the dependence of permittivity and refractive index of an ionized medium upon its electron density and frequency of the passing wave. Appleton [15] has formulated the dependence of the above relations upon externally imposed magnetic fields and has accounted for the loss effects due to the presence of dissipative factors in the medium. Similarly, expressions for intrinsic impedance, reflectivity, and conductivity have been formulated by Everhard and Brown [16], Eccles [17], Baker and Rice [12], and Margenau [18]. These relations have, in turn, yielded expressions for the propagation constant and its dependence upon wave frequency, plasma frequency, and collision frequency. The limitations of Margenau's work to field strengths of low values have been removed and his work extended to some extent by Everhart and Brown [16], and Cahn [19], resulting in relatively simple expressions for conductivity and propagation constant over a wide range of values of field strength, wave frequency, and air pressure.

Approximate Formulations

It is to be noted that much of the work referred to above is based upon investigation of the ionosphere. A great deal of this is predicated on simple classical theory which considers the behavior of average, individual electrons while neglecting their over-all velocity distribution. Generally, though not always, a plane wave is considered and the ionized medium it traverses is assumed to be isotropic, homogenous, constant, and linear. For specific application to ionospheric studies, a slowly-varying stratified distribution of electron density in a uniform magnetic field is often assumed. Further common assumptions consider the energy added to the electrons by the passing electromagnetic wave as small compared to their kinetic energy, effects of electron collisions are neglected, and the associated positive ions of the plasma are assumed immobile [20]. These assumptions generally do not reflect accurately the conditions accompanying propagation of RF signals to and from a vehicle surrounded by a hypersonically-generated ionized sheath, but they have been necessary in order to make possible the mathematical description of even the simpler phenomena and to evolve associated quantitative formulas. Yet, if the results are considered in proper context with full appreciation of the limitations involved, they can be of value in understanding some of the basic phenomena underlying wave propagation through hypersonically generated plasmas.

In order to arrive at such an understanding, it is instructive to consider from a qualitative aspect the electric field of an electromagnetic wave incident upon an ionized region and exerting a force upon the electrons of that region. (The positive ions, being many times heavier than the electrons, are assumed immobile; hence only the electrons are effective in interacting with the electromagnetic wave [21].) Some of the excited electrons collide with heavier gas molecules, giving up part of their kinetic energy to the molecules, and re-radiating the remainder in the form of a distorted radio wave; the result is an absorption of energy from the incident wave. Discounting these occasional collisions with heavier ions and the influence on the excited electrons of either the earth's natural magnetic field or of an artificially imposed magnetic field, the direction of the induced electron vibration parallels the electric field vector of the radio wave. The vibrating electrons behave as miniature parasitic antennas, alternately absorbing and re-radiating the radio-wave energy incident upon the ionized gas. The net effect of this mechanism is an alteration of the phase velocity of the wave, a change in the intrinsic impedance of the medium, and a refractive action wherein the wave path is bent away from the regions of high electron density toward regions of lower electron density [22]. Consider the refractive index; using the model described above, it can be expressed simply as [12]

$$n = \sqrt{1 - \frac{Ne^2}{\epsilon_0 m \omega^2}}, \tag{1}$$

where

e = electron charge in coulombs,
m = electron mass in kilograms,
ω = frequency of wave propagation in radians/second $= 2\pi f$,
N = number of electrons/meter^3, and
ϵ_0 = permittivity of free space.

This equation is derived from the assumption that the average time between collisions of an electron with other particles is appreciably greater than the period of the exciting or impinging wave, *i.e.*, collision frequency ν is less than wave frequency f.

It should be noted in (1) that $n=0$ when $Ne^2/\epsilon_0 m = \omega^2$. This particular value of ω is termed the critical or plasma frequency ω_{cr} and is defined precisely as that propagation frequency at which the refractive index is zero in the absence of a magnetic field. For $\omega > \omega_{cr}$, n is real, and unrestricted wave propagation is present; for $\omega < \omega_{cr}$, n is imaginary and only evanescent waves exist. For $\omega = \omega_{cr}$, complete reflection occurs, and there is no wave propagation through the medium. Thus, by considering the simple model of interacting electric wave vectors and plasma electrons, we have arrived at the concept of critical or plasma frequency, and have gained a knowledge of frequency limits of propagation and their dependence on ionization density.

By using similar techniques based upon simple models, relations for wave attenuation, phase shift, plasma conductivity, intrinsic impedance, and reflectivity have been derived by a number of investigators [12]–[19]. These expressions have been compiled in part by DeVale, *et al.* [23], and Sisco and Fiskin [24]; these are combined below in a common set of units for convenient reference.

attenuation constant
$$\alpha = \omega\sqrt{\frac{\mu\epsilon}{2}\left(\sqrt{1+\frac{\sigma^2}{\omega^2\epsilon^2}}-1\right)} \quad (2a)$$

phase shift constant
$$\beta = \omega\sqrt{\frac{\mu\epsilon}{2}\left(\sqrt{1+\frac{\sigma^2}{\omega^2\epsilon^2}}+1\right)} \quad (2b)$$

plasma frequency
$$\omega_{cr} = \sqrt{\frac{Ne^2}{\epsilon_0 m}} \quad (2c)$$

conductivity
$$\sigma = \epsilon_0\,\frac{\omega_{cr}{}^2\nu}{(\nu^2+\omega^2)} = \frac{Ne^2\nu}{m(\nu^2+\omega^2)} \quad (2d)$$

permittivity
$$\epsilon = \epsilon_0\left(1-\frac{\omega_{cr}{}^2}{(\nu^2+\omega^2)}\right) = \epsilon_0\left(1-\frac{Ne^2}{\epsilon_0 m(\nu^2+\omega^2)}\right) = \epsilon_0 - \sigma/\nu$$

reflection coefficient
$$\rho = \frac{\eta-\eta_0}{\eta+\eta_0}$$

intrinsic impedance
$$\eta = \sqrt{\frac{\mu}{\epsilon\left(1+\frac{\sigma}{i\omega\epsilon}\right)}}$$

index of refraction
$$n = \sqrt{1-\frac{Ne^2}{\epsilon_0 m\omega^2}}$$

relative permittivity
$$\epsilon_r = \epsilon/\epsilon_0 = 1-\frac{1}{(\nu/\omega_{cr})^2+(\omega/\omega_{cr})^2}$$

relative permeability
$$\mu_r = \mu/\mu_0 = 1$$

gyromagnetic frequency
$$f_h = \frac{e\mu_0 H}{2\pi m}.$$

In the above expressions:

e = electron charge in coulombs,
m = electron mass in kilograms,
N = electron density (electrons/meter^3),
ϵ = permittivity of plasma,
ϵ_0 = permittivity of free space,
μ = permeability of plasma,
μ_0 = permeability of free space,
ϵ_r = relative permittivity,
n = refractive index,
η = characteristic impedance,
γ = propagation constant $= \alpha + j\beta$,
η_0 = intrinsic impedance of free space,
ν = collision frequency,
H = magnetic field (ampere turns/meter).

In order to demonstrate the nature of the dependence of some of these parameters on electron density (and, therefore, on vehicle velocity and plasma frequency), the attenuation and phase shift have been plotted in Figs. 3 and 4 over a wide range of propagation frequencies. Note in Fig. 3 the large attenuation which accompanies high electron densities (plasma frequencies), and the rapidity with which attenuation drops off as wave frequency approaches plasma frequency. It is interesting to note that the frequency range of greatest attenuation (100–1000 mc) in Fig. 3 corresponds to a region approximating phase shifts which are comparable to free space values (see Fig. 4). In other words, if one desires to obtain normal phase shifts in propagating through a plasma, one must then contend with high

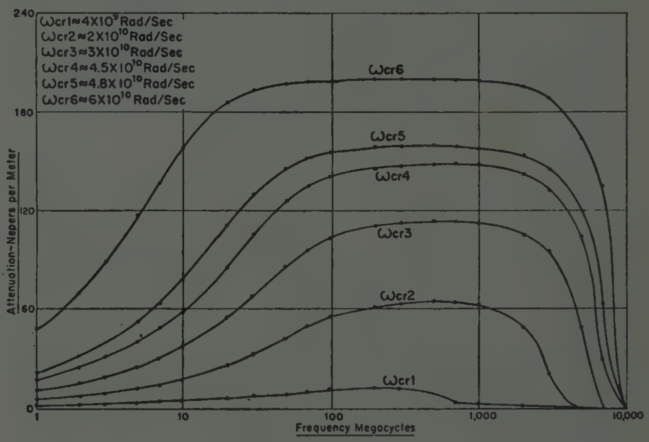


Fig. 3—Attenuation vs frequency for various plasma frequencies (electron densities).

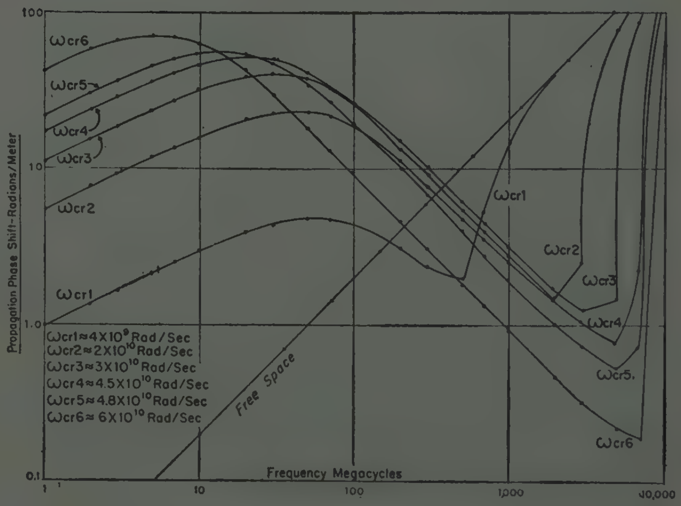


Fig. 4—Propagation phase shift vs wave frequency for various ionization levels (plasma frequencies and electron densities).

wave attenuation. Note also in Fig. 4 that as wave frequency approaches plasma frequency, the phase constant undergoes violent variations. This points up the desirability of utilizing the lowest of wave frequencies for propagation; as shall be noted later, however, low wave frequencies introduce unique problem areas.

Since wave attenuation and phase shift are such important parameters in radio communications and telemetry, let us now consider the influence of a superimposed magnetic field upon the attenuation and phase of a wave passing through a plasma. Again, this can be explained in terms of our oscillatory electron and wave vector model. It has been found that at high wave propagation frequencies the velocity of vibration of the excited plasma electrons is practically unaffected by application of a mild external field, *i.e.*, the excited vibrating electrons are placed into paths which are only slightly elliptical, deviating but little from the linear motion described earlier. Accordingly, the effective conductivity of the plasma is practically equivalent to that of the zero-magnetic field case of (2d). As the wave frequency is lowered, the vibrational velocity of the excited electrons increases and their paths become elliptical and then circular [25]. At a frequency value expressible as [26]

$$\omega_h = \frac{e\mu_0 H}{m} = 2\pi f_h, \tag{3}$$

the electron vibrational velocity is greatly increased, and the conductivity of the plasma becomes correspondingly great. This, of course, is manifested by an increase in wave attenuation. The particular frequency f_h at which these phenomena are observed is termed the gyromagnetic frequency of the medium. At frequencies greatly less than the gyromagnetic frequency, the applied magnetic field serves to decrease the electron vibrational velocity well below the value which would be present were the magnetic field absent. Thus, electron collisions and, consequently, plasma conductivity is decreased, resulting in decreased attenuation of passing electromagnetic waves. This behavior has been confirmed for high-frequency radio waves propagating through the earth's atmosphere with the earth's magnetic field acting as the influencing magnetomotive force. The earth's magnetic field has a value of approximately ½ gauss; using this value in (3) gives a value of 1.4 mc for the electron gyromagnetic frequency. This corresponds to a wave propagation wavelength of 214 meters. Nichols and Schelling [25] have found that marked wave attenuation of propagated waves occurs at precisely this frequency, with a decrease of attenuation taking place at longer and shorter wavelengths. This behavior is also recorded by Baker and Rice [12].

In view of this influence of the earth's natural magnetic field upon wave attenuation and phase shift, one is led to conjecture the effects of artificially imposed magnetic fields on wave propagation. This point is considered in the following section in which the possibility of this approach is considered.

An indication of the dependence of plasma frequency upon electron density is presented in Fig. 5. Calculations have shown that the electron densities generated by a hypersonic aerodynamic vehicle lie close to the curve of Fig. 5 if the axis of abscissas is calibrated in Mach number (see top of Fig. 5). This near-coincidence of curves remains valid up to about Mach 14 or 15, after which other processes set in, which cause the vehicle velocity vs plasma frequency relationship to deviate from the electron density vs plasma frequency curve of Fig. 5. This frequency-velocity-electron density relationship means that RF telemetry or communication can be carried on to and from the re-entering hypersonic vehicle only at frequencies equal to or greater than the critical or plasma frequency corresponding to the particular re-entry velocity. Thus, at Mach 10, the frequency must be above 200 or 300 mc; at Mach 17 the minimum usable frequency is 5000 mc, etc. Frequencies immediately below the plasma frequency will not penetrate the ionized sheath surrounding the re-entry vehicle, and hence are unsuitable for establishment of telemetry or communications. This is particularly disconcerting in present day telemetry and communication practice, due to the extremely heavy equipment investment in the VHF-UHF range. In view of this, the importance of devising means of transmittal through the hypersonically-generated plasma of VHF-UHF radio waves is easily appreciated.

Alternate approaches utilizing microwave frequencies

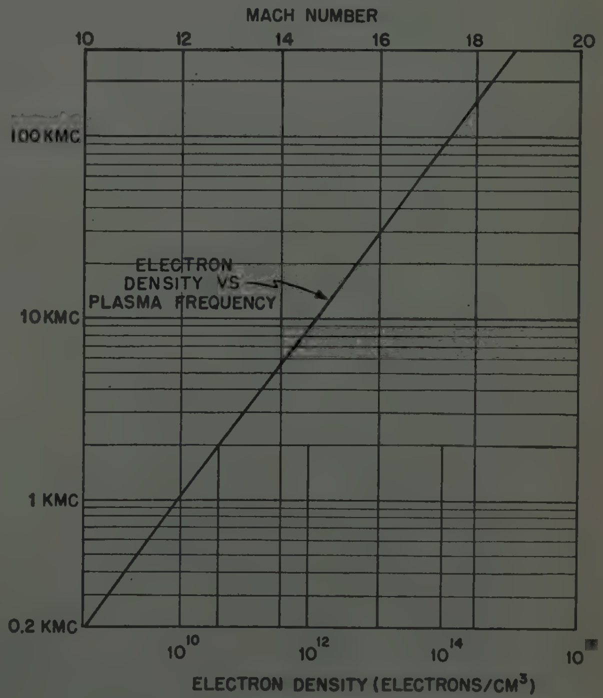


Fig. 5—Plasma frequency as a function of electron density and Mach number.

on the one hand or very low radio frequencies on the other, though offering some solutions to the propagation problem, introduce their own peculiar complications. These will be discussed below.

Exact Formulations

The task of devising, analyzing, and evaluating the exact mechanisms of electromagnetic wave and plasma interaction at the lower radio frequencies is extremely difficult due to the complicated nature of the plasma itself, the proximity of vehicle antennas to the plasma, and the difficulty of reasonably expressing the interactions and relations existing between the two. The dimensions of the shock front and of the various layers constituting the ionic sheath are not constant, but vary with vehicle velocity and environmental conditions—and consequently with temperature and density. The composition of these regions varies also in space and time, thus making untenable the convenient concepts of homogeneity, isotropy, and constancy so effectively used in common wave propagation analyses. Because the antenna is either embedded, in contact with, or very close to the plasma sheath, the effect on the near field radiation components is substantial. As a result the input impedance characteristics are different from the free-space radiation case. Furthermore, since the plasma sheath thickness is generally small, or comparable to the antenna dimensions, the fields arriving at its outer boundary do not constitute a plane wave; thus the conventional coefficients of transition, *viz.*, reflection and transmission, do not apply, and the over-all analysis is accordingly made more complex.

Determination of the nature of the plasma itself is complicated by the gas behavior at the high temperatures associated with hypersonic velocities. Under these conditions, energy is consumed in exciting the translational, rotational, and vibrational degrees of freedom of the gas molecules, the energy levels of atomic electrons are increased, and outer orbital electrons are liberated from their atoms and made available to react with incident electromagnetic waves. Also erosion, sputtering, and mass ablation from the vehicle surface introduce impurities into the ionic cloud which further complicate the description of wave and particle interaction.

Because of the complexity of these processes, and the present lack of knowledge concerning their over-all qualitative and quantitative nature, exact analyses and formulations are prohibitively difficult. Some excellent work has been started in this regard [24], [27], [28], but much of it deals only with specific (and sometimes isolated) aspects of the over-all problem. The truly comprehensive approach must take into account the specific nature and behavior of the medium in all of its varied interactions with electromagnetic waves. Since much of this knowledge cannot be adequately arrived at on purely theoretical grounds, recourse must be had to experimental processes such as those described by Hertzberg [11] and Yoler [29]. In this manner, the validity or falsity of certain assumptions employed in the theory can be ascertained with resulting mutual benefit to both disciplines. It is expected that this would be an iterative process resulting in specific mathematical models which are continually readjusted, appraised, and formulated in the light of confirming and guiding experimentation. When the theoretical formulations have progressed to the point where the potentialities of laboratory experimentation are exhausted, the process would then be advanced to actual flight testing in vehicles which are even now in the design and advanced development stage [30], [31]. It is emphasized that in attacking a problem of this complexity, isolated approaches on specific phases, be they experimental or theoretical, are insufficient to produce the breakthrough that is demanded. This can be realized only by a unified and integrated concentration of many disciplines brought to bear simultaneously upon the common objective.

NEW APPROACHES FOR ACHIEVING TRANSMISSION OF UHF-VHF SIGNALS THROUGH HYPERSONICALLY GENERATED PLASMAS

It has been stated earlier that transmission of RF signals through the hypersonically generated plasma can be effected by utilizing extremely high frequencies such as microwaves (and even optical frequencies) [32] or, on the other hand, by using low radio frequencies such as VLF. The attenuation and phase shift curves of Figs. 3 and 4 and the discussions accompanying them indicate the advantages of using frequencies at either end of the RF band. Let us now look at these viewpoints more closely. Consider first the utilization of microwave or higher frequencies. First, it is evident that since most of the telemetry, guidance, and communication equipment presently in use is designed mainly for the VHF-UHF band, an extensive investment in equipment and technique development would be required before these higher frequencies could be effectively employed for the extensive uses mentioned above. Secondly, interaction of microwaves with the dipole moments of the particles formed in the hypersonically generated plasma are to be expected under certain conditions. Atmospheric attenuation processes are also present which superimpose on the usual inverse square range dependence an exponential decrease of wave intensity with distance from the source [33]. In view of these characteristics, the advantages of microwaves over lower frequencies are few for the applications considered here.

Now consider the VLF range. Again, since most of the present-day telemetry and airborne communication equipment is not in this band, the major drawback of extensive equipment development in the VLF range is present. Design of such equipment for airborne use is fraught with many problems because of the long wavelengths and correspondingly large size of the associated components which are required. As a result, heavy penalties are to be paid in terms of weight, size, and power consumption if VLF equipment is considered for airborne application. Further, since the ionosphere is essentially reflective to VLF waves, propagation problems are to be expected if portions of the vehicle flight profile lie within the various ionospheric layers of the earth [34]. Thus, as a greater number of details concerning the nature and equipment of VLF, microwave, and optical frequencies are considered, the more evident becomes their unsuitability for solving (except perhaps as a last resort) the problem at hand.

In view of this, the desirability of propagating telemetry, guidance, and communication signals within the established VHF-UHF bands is evident. Since, however, the plasma is essentially opaque to these frequencies, it becomes necessary to devise new approaches whereby the transmissivity of the plasma may be increased for signals in these bands. It appears that most such approaches can be grouped into two categories: 1) those which depend upon removal of the ionic sheath or in some way altering its deleterious influence on RF waves, and 2) those which seek to advantageously utilize the presence of the sheath by introducing modifications which will permit the sheath to actually enhance RF propagation. The former are conveniently referred to as recombination and cooling techniques; the latter as electric and magnetic interaction methods. These will now be considered in the above-named order.

Recombination and Cooling Techniques

The techniques of this first category seek to remove the electron influence on electromagnetic fields by removing the electrons themselves through such processes as diffusion, chemical interaction, and application of static electric and magnetic fields to the ionized regions. Two such suggested approaches depend on local cooling of the vehicle and chemical neutralization of the plasma [35], [36]. These methods are of interest not only to the electrical engineer concerned with propagation of RF signals through the sheath, but also to the aerodynamicist concerned with the heat absorption problem of the re-entrant vehicle. The referenced local cooling methods are based on separation of the laminar air flow over a hypersonic vehicle. This is induced by means of a "spike" in the vehicle's nose and "razor blade" extensions of all leading edge surfaces. The resulting detached air flow [37] lowers the equilibrium temperatures of the vehicle surfaces, and injection of a gas into the resulting separation space can theoretically reduce the heat transfer to the body. Such a technique serves to decrease the electrical conductivity of the ionic sheath (and therefore increases the transmissivity of radio waves) by eliminating the contamination products of ablation and vehicle surface sputtering. The problem of the dense electron concentration within the plasma itself still remains, however.

Cooling of the region surrounding a hypersonic body may be effected by specific endothermic chemical reactions such as, for example, the formation of acetylene and monatomic hydrogen from diatomic hydrogen with a total heat absorption of approximately 100 calories per gram molecule [36]. The problem here lies in the development of stable compounds which are highly endothermic in order to reduce the quantity of chemicals which must be transported in the vehicle and circulated over its surfaces. A related process for decreasing ion concentration within the plasma depends upon the mechanism of diffusion, *i.e.*, the movement of ions solely on the basis of their thermal motion. The comparative thinness of the boundary layer surrounding a hypersonic vehicle, and the large temperature differences existing on either side of it, indicate the presence of large thermal gradients within that layer. These thermal differences are instrumental in attracting atoms of the hot gases of the shock layer to the surface of the vehicle where they can recombine to give up their heat of dissociation. Increase of this thermal gradient by cooling of the vehicle surface can conceivably result in larger scale diffusion processes and consequently greater degrees of deionization. It is in this respect that the dissociation mechanism is related to the above-mentioned cooling processes.

Suggestions dealing with chemical alteration of the plasma seek to inject substances into the plasma which have an affinity for electrons. Thus the deleterious interactions of the electrons with RF waves are removed by attaching the electrons to heavier ions which are less mobile than electrons and therefore less capable of interacting with the passing waves. Presumably, such substances would be those whose outer atomic shells are short of completion by one or two electrons and would therefore readily accept free electrons. Thus, the outer, or L, shell of fluorine is composed of seven electrons—just one short of the full complement of eight allowable L electrons. Presumably, then, fluorine would be a good acceptor of electrons. Similarly, oxygen, which lacks two electrons in its outer shell, would also be a good acceptor. There evidently exists a wide variety of substances which are well-suited to act as electron acceptors. It remains to determine the nature of their interactions at the high temperatures involved in a hypersonically generated plasma and the nature of the ions and compounds which are formed in these interactions. The accomplishment of this must depend in great part on experimental measurements in high temperature facilities such as shock tubes and plasma torches.

In a similar vein, related suggestions have been put forth concerning use of artificially generated magnetic fields upon the plasma in order to decrease RF attenuation and generally improve wave transmission through ionized regions. The prevalent mechanism here has been mentioned in the previous section. It was stated there that at frequencies much lower than the gyromagnetic frequency, applied magnetic fields will act to decrease the electron vibrational velocity far below the value with which it would oscillate in the absence of the fields. This is manifested as a decrease in effective plasma conductivity and, consequently, a decrease in the wave attenuation properties of the medium. From (4) gyromagnetic frequency may be expressed as $f_h = 3.52 \times 10^4 H$, where H represents the applied magnetic field in ampere turns per meter. Assume that a field of 500 gauss could be generated through a particular region of the plasma. In this case, f_h is approximately equal to 1400 mc. Radio frequencies in the VHF communication and telemetry band lie well below this value. Thus, such frequencies are capable of decreasing the plasma conductivity as mentioned earlier and are potentially capable of creating a transmission window through the hypersonically generated plasma. It is recognized that generation of large magnitude magnetic fields in an airborne vehicle and through an associated plasma of even moderate physical dimensions is extraordinarily difficult; hence if such a technique is not to remain merely an academic curiosity a great deal of effort must be devoted toward practical means of establishing and maintaining such a field.

Electric and Magnetic Interaction

Among the methods of this second category which seek to utilize the plasma sheath for enhancement of RF transmission, two stand out as offering some promise [38]. These describe conditions which influence the transmission of electromagnetic radiation through ionized regions. The first of these, based on the magnetoionic theory [15], [39], is concerned with parameters which either permit or restrict transmission. The second, which has its foundation in the electromagnetoionic theory [40], postulates actual wave amplification during propagation through ionized regions when certain conditions are fulfilled.

Consider first the magnetoionic postulates. The propagation of a plane electromagnetic wave in a homogenous ionized gas depends on the electron density N, the mean frequency ν of collision of an electron with a molecule or ion, the electromagnetic wave frequency f, the steady magnetic field H, and the angle between H and the direction of propagation. The magnetic field can be expressed in terms of the gyromagnetic frequency f_h of a free electron; electron density is commonly expressed in terms of the plasma frequency f_{cr}. In describing the phenomena of interest, it is common to utilize three parameters which are expressed as follows:

$x = (f_{cr}/f)^2$ which is proportional to N,

$y = f_h/f$ which is proportional to H,

$z = \nu/2\pi f$ which is proportional to ν.

In the absence of magnetic fields, and for small electron collision frequency, the refractive index of the medium may be expressed as

$$n^2 = 1 - x.$$

This is recognized as (1). Normal propagation exists when x is less than unity. When $x = 1$, the refractive index becomes zero, which indicates a condition of complete wave reflection; for $x > 1$, the refractive index becomes imaginary. Physically, this is indicative of extreme attenuation of the wave within the medium, a condition analogous to the phenomenon of attenuation in a waveguide beyond cutoff. For $x < 1$, propagation within the medium is permissible, subject to attenuation due to collision of excited electrons with ions. This attenuation is $\alpha = \nu x/nc$ where c is the velocity of light. As x is increased from its less than unity value, which is caused by increasing the number of electrons in the region, ν increases and n decreases; thus the refractive index rapidly approaches zero. When a steady magnetic field is superimposed upon the ionized region, the electromagnetic ray traversing the region divides into two modes, the so-called ordinary and extraordinary rays. (Piddington [42] has shown that the magnetoionic medium is really quadruply refracting and not merely doubly refracting. This conclusion is reached when heavy-ion motions are taken into account in the analysis—a fact which we find unnecessary to consider in this paper.) Depending upon the magnitude and orientation of the magnetic field, the emerging waves are elliptically, circularly, or linearly polarized. In considering the passage of electromagnetic radiation through the ionized region, superposition of a static magnetic field of varying intensity throughout the region can lead to conditions conducive to escape of the wave [26]. Fig. 6 shows a plot of x (proportional to electron density) vs y (proportional to steady magnetic field). The points A, B, and C represent the values of x and y at corresponding points in the ionized region. An electromagnetic wave passing out of the plasma is represented by the ABC path approaching the origin of the xy-coordinate system. When the wave emerges from the plasma, the point O of Fig. 6 is reached. Thus, if x and y (*i.e.*, electron density and magnetic field intensity) can be made to decrease in proper magnitude as the wave emerges to the edge of the ionized region, then escape of the wave is possible. Curves II and III denote other possible modes of electron and magnetic field decay.

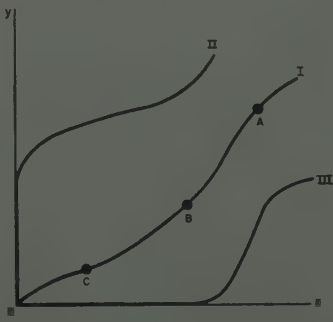


Fig. 6—Representative modes of decay of electron density and magnetic field (after Pawsey and Bracewell). $x \propto$ electron density; $y \propto$ magnetic field intensity.

A further analysis of these magnetoionic mechanisms indicates that definite pass and stop regions exist in a plasma which has a magnetic field superimposed on it. These regions are separated by surfaces expressible as zeros or infinities of the refractive index. Zeros occur at $x=1\pm y$ for longitudinal propagation (direction of wave travel parallel to magnetic field) and for $x=1$ and $x=1\pm y$ for transverse propagation (direction of wave travel perpendicular to magnetic field). Infinities occur for

$$y=\sqrt{\frac{1-x}{1-x\cos^2\theta}}$$

where θ denotes the angle between magnetic field and direction of wave propagation.

The zeros and infinities for the longitudinal case are depicted in Fig. 7 for the ordinary and extraordinary rays; the corresponding stop and pass regions are shown in Fig. 8(a) and 8(b). It is seen that the introduction of a suitable magnetic field, combined with a controlled electron density, permits propagation of waves in certain regions of high electron density which, without magnetic fields, would be stop regions.

It is important to note, however, that the magnetoionic theory is strictly applicable to radio astronomy phenomena. Thus, it is predicated upon conditions involving shallow gradients of electron density and magnetic field. It is doubtful that the electron distribution in hypersonically generated plasmas conforms to this requirement, although it is conceivable that some suitable alteration in electron distribution could be effected. If this can be accomplished together with generation of suitable magnetic fields, it is probable that the magnetoionic approach may offer some promise in solving the re-entry telemetry and communication problem.

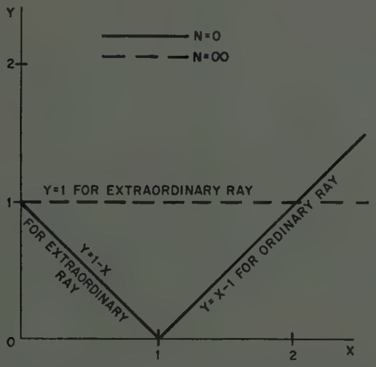
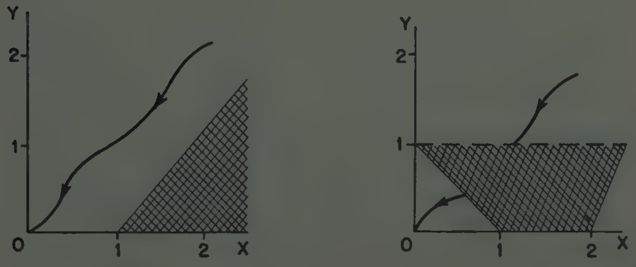


Fig. 7—Zeroes and infinities of refractive index for the magneto-ionic modes (after Pawsey and Bracewell). $x \propto$ electron density; $y \propto$ magnetic field intensity.



(a) (b)

Fig. 8—Stop (shaded) and pass regions for the magneto-ionic modes (after Pawsey and Bracewell). (a) Ordinary ray. (b) Extraordinary ray. $x \propto$ electron density; $y \propto$ magnetic field intensity.

Consider now the electromagnetoionic theory which postulates electromagnetic wave amplification during propagation through ionized regions when certain conditions are fulfilled. This theory is a generalized description of the behavior of plane waves in an ionized medium pervaded by static magnetic and electric fields; as a limiting case it includes the well-known magnetoionic theory, and its application is subject to the same restrictions governing the magnetoionic theory. It is developed from the following laws of physics:

1) Maxwell's laws of the electromagnetic field.
2) Conservation of electrons and positive ions.
3) Maxwell's laws of the transfer of momentum in mixtures of different kinds of particles which are subject to fields of force.

These laws can be expressed completely by a set of six equations relating electric and magnetic fields, scalar and vector potentials, electron drift velocity, and induced perturbation effects. These are reducible to a set of three simultaneous equations whose solution yields the equation of dispersion; from this equation, the varying electric and magnetic field vectors are derived.

When the dispersion equation is plotted as a function of wave frequency ω, the resulting curve is found to cut the ω axis at specific points which define the limits of frequency bands within which the refractive index of the medium is purely imaginary. In accordance with magnetoionic theory, these are regions of extreme wave attenuation—essentially stop regions for electromagnetic propagation. Pursuing the matter further, Bailey [40] has shown that when an electron drift velocity exists, the corresponding curve has some branches similar to the ones considered above, but distorted in a skew manner so that they become unsymmetrical about the ω axis. As a result, bands are formed in which the refractive index is not purely imaginary, but a complex number. This indicates that the effect of electron drift creates wave amplification in frequency bands in which otherwise waves cannot be propagated. A further implication of the theory is that no wave amplification is possible when the wave is propagated transversely to both the drift velocity and the applied steady magnetic field. However, when the electron drift and magnetic field have a common direction, wave amplification is increasingly favored by orientation of the direction of propagation toward this common direction.

Once again it must be emphasized that the tenets of this theory, like those of the magnetoionic theory, are based upon radio astronomy phenomena. Hence, prior to their application to problems of electromagnetic transmission through hypersonically-generated plasmas, it is essential to determine the relative magnitude of the effects involved and the scaling relations which may need to be employed in order to achieve workable relations among the various phenomena involved. Thus it is false to assume that these theories can be applied in blanket fashion to the electromagnetic propagation problems outlined in this paper. What is implied, however, is that these theories appear capable of providing the essential analogous elements whereby adjustments and alteration of the physical conditions imposed by plasmas and waves can be effected to produce constructive rather than destructive interference in their mutual interaction.

DISCUSSION

Application of the classical ionospheric treatments of electron-ion-radio wave interactions to the plasma-radio wave interference problem can yield only limited workable results. This is attributable to the fact that the plasma is more complex than the ionosphere, sharply varying, and composed of small regions with relatively sharp boundaries. The theoretical treatment which is required to adequately describe the phenomena involved is extremely complicated, and probably cannot be accomplished without the aid of associated laboratory and free flight experimentation.

Although many of the difficulties associated with RF transmission through plasmas can be removed by use of very low or very high (microwave) radio frequencies, this recourse leads into economic problems due to the heavy equipment investment in the VHF-UHF telemetry and communication bands. Further, the high frequencies in the microwave region are susceptible to their own peculiar interactions with the molecules of the plasma and with the atmosphere itself. The very low frequencies, on the other hand, extract heavy penalties in terms of size and weight of the components which constitute the equipment, and present the added disadvantage of ionospheric reflection—a serious drawback when the re-entry vehicle is above or within the various layers of the ionosphere. This leads to the conclusion that intensive effort should be devoted to establishing means of transmitting VHF-UHF signals through the hypersonically-generated plasma sheath, and that microwaves or very low frequencies should be employed only as a last resort.

Suggested techniques for propagating VHF-UHF signals through the plasma depend upon recombination processes (to remove the plasma electrons and their deleterious influence) or imposition of auxiliary electric and magnetic fields on the electrons and ions (to induce passage and/or amplification of the passing RF wave). Some theoretical difficulties are to be anticipated in applying the latter approach to the present problem because it has its roots in radio astronomy phenomena where the magnitude of the magnetic fields, electron drifts, and ion distributions differ from those of the hypersonically generated plasma. However, it is possible that judicious alteration of some of the plasma parameters, together with suitable adjustment of associated boundary conditions, can introduce sufficient degrees of similarity to permit application of the electromagnetoionic concepts.

Some experimental difficulties are to be anticipated in application of both the recombination approach and the electromagnetoionic approaches because of the magnitude of the auxiliary electric and magnetic fields which must be employed in order to bring about the desired interactions. However, these are not insurmountable, and proper alteration of assorted plasma parameters may serve to diminish the severity of the imposed field requirements.

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An Analog and Digital Airborne Data Acquisition System*

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***Summary*—A highly flexible analog and digital airborne data acquisition system has been designed and developed by Radiation, Inc., for the General Electric Company Flight Test Facility at Edwards Air Force Base, Calif.**

The acquired data is recorded on one-inch tape. The format may be fourteen tracks of analog or seven tracks of analog with sixteen tracks of digital data. Ampex AR-200 equipment is used on the analog subsystem. Complete analog system reference data are recorded on one track.

The digital system has extreme flexibility. From ten to ninety channels may be programmed for individual word rates of from 1.25 samples per second to 2×10^3 samples per second. System word rates are 20 kc, 10 kc, 5 kc, and 2.5 kc. The analog-to-digital converter performs up to 2×10^4 conversions per second, depending on the selected sampling rate of the multiplexer. Time-history and system marker data are also recorded. Provision for acceptance of Gray code digital input data has been made. The input on any channel may be low-level (±10 mv full scale) or high-level (±5 volts full scale). Signal conditioning and automatic calibration are provided. Multiplexer channel capacity can be extended by the addition of more modules. Flexibility of programming is provided with pluggable program boards.

The system is designed using all transistorized components and is constructed to withstand vehicular environmental extremes. Detailed descriptions of the low-level gate circuit, programming, accumulating, and coding techniques are made.

INTRODUCTION

THE introduction of digital techniques to the field of flight testing has offered possibilities of increased flexibility, capacity and accuracy to the flight test engineer. The initial development of the first airborne pulse code modulation (PCM) telemetry was initiated with the development of the AKT-14 for WADC in 1954. Since then, the evaluation of solid state component development and circuit design has led to highly reliable, small, and light-weight data gathering systems. The ruggedness and stability of a system of this type has been proven by the success of the PCM telemeter in use on the rocket sleds at Holloman Air Force Base, New Mexico. The advantages of digital data handling are gained when transmitting over a telemetry link or recording in the vehicle. The accuracy of

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the digital data is easily maintained with the application of either media.

This paper is an attempt to show the unique circuits, logic, and packaging techniques employed by Radiation, Inc., Melbourne, Fla., to meet the in-flight data acquisition system requirements outlined by B. E. Applegate.[1] The system was designed by the General Electric Jet Engine Flight Test Facility at Edwards Air Force Base, Calif. Results of design and development for the digital subsystem are described in the areas of analog signal conditioning and calibration, time-division low- and high-level multiplexing, programming of the electronic multiplexer gates, high resolution amplification, coding, data handling, and recording. The use of magnetic tape for analog and digital data correlation is made possible by the development of a unique digital head by Ampex Corp., Redwood City, Calif. The analog subsystem is primarily a result of development by Ampex and is described in detail by E. P. Brandeis and M. E. Harrison.[2]

SYSTEM DESCRIPTION

The over-all data acquisition system consists of both analog and digital subsystems. Fig. 1 indicates the major blocks that are required to form the complete system. Subsystem *A* (Analog) is above the dotted dividing line and subsystem *D* (Digital) is composed of those blocks below the line. Since the components of the analog subsystem (subsystem *A*) are adequately described in a previous technical paper,[2] no details of the components will be furnished except where required for a clear understanding of the over-all system operation. Each component of the digital subsystem (subsystem *D*) will be described in varying detail, the extent of which will depend on the uniqueness of circuit design, logic, or packaging techniques conceived in the development of that component. the following are the major specifications of the digital subsystem:

1) *Electrical Specifications*

Over-all System Accuracy.........	±0.2 per cent referred to full scale input
Input Signal Levels..............	±10 mv, ±5 volts
Word Rates......................	20 kc, 10 kc, 5 kc, 2.5 kc
Sampling Rate Range	2×10^3 to 1.25 samples per second
Programming....................	Variable

2) *Environmental Specifications*

Temperature.....................	−55°C to +71°C
Altitude.........................	100,000 feet
Vibration.......................	5 to 50 cps @0.060 inch (*DA*) 50 to 500 cps @0.010 inch (*DA*), 10g maximum
Shock..........................	10g in all directions
Power..........................	400 cps ±20 per cent −115 volts ±20 per cent

The primary considerations in the design of the signal conditioning and calibration circuits were flexibility, simplicity and reliability. No active components are used since only attenuation of input signals is required. Flexibility is obtained by an arrangement of terminals on a card which allows the interconnection of points within the circuits. The configuration of the wiring may be tailored to accept one of the three basic types of measuring devices: 1) bridge type transducer, 2) thermocouple, and 3) potentiometer. Adjustment of resistor values may be made, if necessary, to bring the signal within the measurable range of the system for the various signal levels expected from wide variations in transducer characteristics and parameter excursions. Calibration signals, also variable, are furnished for

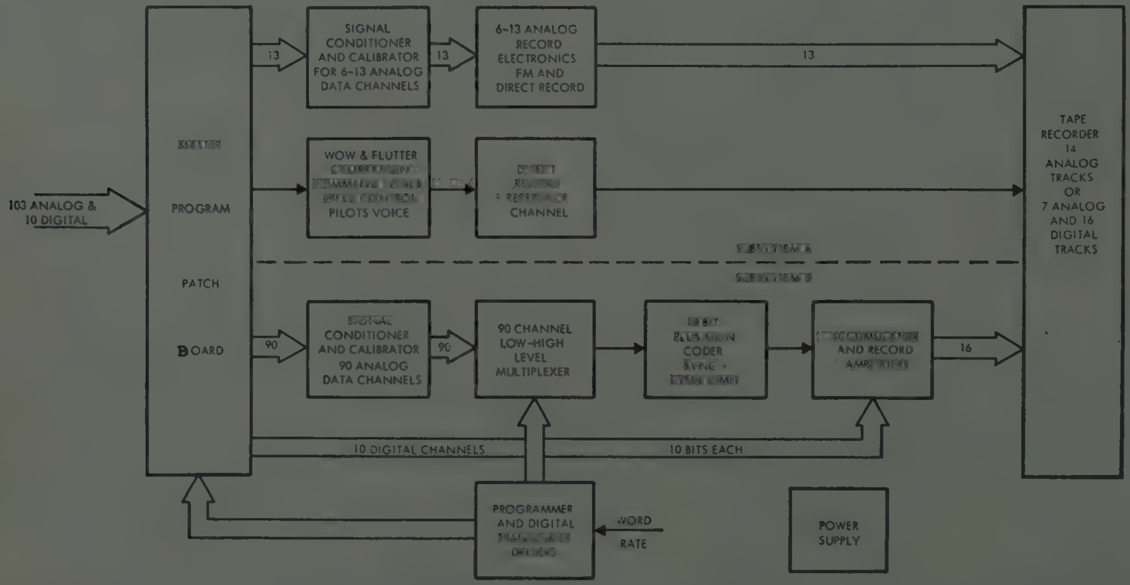
103 ANALOG &
10 DIGITAL
MASTER
PROGRAM
PATCH
BOARD
13
SIGNAL CONDITIONER AND CALIBRATOR FOR 6-13 ANALOG DATA CHANNELS
13
6-13 ANALOG RECORD ELECTRONICS FM AND DIRECT RECORD
13
WOW & FLUTTER CALIBRATION COMMAND TONES SPEED CONTROL PILOTS VOICE
DIRECT RECORD + REFERENCE CHANNEL
TAPE RECORDER 14 ANALOG TRACKS OR 7 ANALOG AND 16 DIGITAL TRACKS
SUBSYSTEM A
SUBSYSTEM D
90
SIGNAL CONDITIONER AND CALIBRATOR 90 ANALOG DATA CHANNELS
90
90 CHANNEL LOW-HIGH LEVEL MULTIPLEXER
10 BIT PLUS SIGN CODER SYNC + CODE LIMIT
ACCUMULATOR AND RECORD AMPLIFIERS
16
10 DIGITAL CHANNELS
10 BITS EACH
PROGRAMMER AND DIGITAL TRANSDUCER DRIVERS
WORD
RATE
POWER SUPPLY

Fig. 1—Analog and digital data acquisition system.

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[1] B. E. Applegate, "A Unique Digital Analog Data Acquisition and Processing System for Flight Testing," presented at the National Symposium of Telemetering, Miami, Fla.; September 23, 1958.

[2] E. P. Brandeis and M. E. Harrison, "Some New Techniques in Airborne Data Acquisition," presented at the National Symposium on Space Electronics and Telemetering, San Francisco, Calif.; September 29, 1959.

functional checking of the entire airborne system and eventual data correction, if required, by the ground data handling system. The calibration signals for the digital subsystem channels are derived from the transducer drive voltage supply and therefore provide a check on that component of the system. Transducer voltages of both polarities are provided for calibration of selected channels in the negative region for a complete system linearity check. Calibration of the analog subsystem channels is required for compensation of the ground recovery components due to drift in the airborne recording equipment. The FM record channels are calibrated at zero and 80 per cent of maximum excursion with voltage from the transducer supply. The direct record channels are calibrated at only full-scale level. The voltage source is a 10.5-kc crystal-controlled oscillator which is also used for a full-scale calibration tone on the reference channel. An 11.7-kc tone is recorded on the reference channel to indicate a zero calibration period. Signal conditioning for these channels is accomplished with circuits identical to those in the digital subsystem.

The transducer voltage source is highly regulated and stabilized over the range of ambient temperature excursion. A unique feature of the supply is that the reference is derived from the coder weighted current reference voltage supply. Excursion of the coder reference tends to be compensated by a related excursion of the transducer supply voltage.

The multiplexing and programming portion of the digital subsystem consists of seven modules. Six of these modules are multiplexers which contain the gates and amplifiers. The seventh module is the programmer. Each multiplexer module contains fifteen gates which are universal in that they may be used for low- or high-level gating. Low-level for this system is defined as the range ±10 mv. High-level is defined as the range ±5 volts. The low-level signals are amplified after multiplexing. There is one amplifier in each multiplexer module which amplifies the ±10-mv signals to the high-level ±5-volt level. The multiplexing and programming system was designed with the greatest flexibility feasible within the practical hardware size limitations.

Fig. 2 is a block diagram of one multiplexer module. Each low-level gate is queried individually from the programmer. Ten digital transducers, such as shaft encoders, may be queried by the programmer instead of any ten multiplexer gates. One type of gate switches either high- or low-level input information. Provision is made to route the low-level signals through the amplifier and route the high-level signals to bypass the amplifier. The amplified low-level signals are again multiplexed onto the final multiplexer buss with the signals originally high-level, and with similar signals from the other five multiplexer modules through a standard gate used in this case as a high-level gate. The gate is electronic and is composed of silicon transistors. The configuration is discussed in some detail later in the paper.

A frame of digital information recorded on the tape contains data calibration signals, range time, and unique word markers in the positions as shown in Fig. 3.

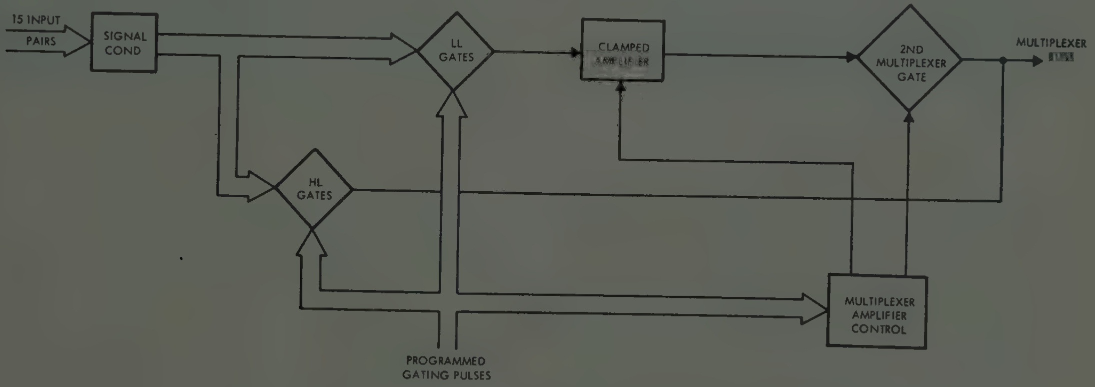
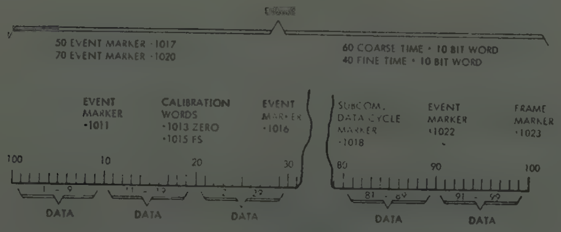

15 INPUT
PAIRS
SIGNAL COND
LL GATES
CLAMPED AMPLIFIER
2ND MULTIPLEXER GATE
MULTIPLEXER BUSS
HL GATES
MULTIPLEXER AMPLIFIER CONTROL
PROGRAMMED GATING PULSES

Fig. 2—One multiplexer (of six). Block diagram.


50 EVENT MARKER ·1017
70 EVENT MARKER ·1020
60 COARSE TIME · 10 BIT WORD
40 FINE TIME · 10 BIT WORD
EVENT MARKER ·1011
CALIBRATION WORDS ·1013 ZERO ·1015 FS
EVENT MARKER ·1016
SUBCOM. DATA CYCLE MARKER ·1018
EVENT MARKER ·1022
FRAME MARKER ·1023
100
10
20
30
80
90
100
1 - 9
11 - 19
22
81 - 89
91 - 99
DATA
DATA
DATA
DATA
DATA

SUPERCOMMUTATED CHANNELS USE EVENLY SPACED TIME POSITIONS IN THE FRAME. I.E., SCC # X MIGHT BE IN TIME POSITIONS 7, 37, 57, 77, & 97

Fig. 3—Frame format.

Plus 5-volt and minus 5-volt transducer power supplies are provided in the system. The current available from each of these power supplies is one ampere and regulation is 0.05 per cent. Provision has been made in the multiplexers for the signal conditioning described above.

The analog-to-digital conversion unit has evolved from many designs for both airborne and ground PCM systems. The unit is completely solid-state, packaged in a volume of about 400 cubic inches. The coder receives the multiplexed, amplified, amplitude-modulated analog sample of about 41 μsec width. Sample and hold circuitry at the input of the coder selects a six-μsec increment near the end of the multiplexed sample, and holds the average amplitude of that sample to within the tolerance required. The output of the analog-to-digital converter is a serial eleven-bit binary code limited to a maximum of 2032 for plus five volts input and a minimum of 0016 for minus five volts input. The maximum word rate is 20 kc, but slower command rates may be selected manually at the programmer. The bit rate of the coder is crystal-controlled at 280 kc.

All data are brought together in the module called the accumulator. System data are generated within this module. The 30-card module performs a total of eleven different functions in one register and peripheral circuits. The module contains the digital recording amplifiers and the time-history generator. Eleven-bit serial code is transformed to ten-bit plus sign, or folded code. A serial-to-parallel conversion is performed. Digital input data are routed into the PCM subsystem directly into the accumulator register. Digital input data are commonly in Gray code form, so a Gray to binary conversion is performed (in the register). Unique words are generated within the accumulator. These are words that mark frame time position, subframe time position, external events, and calibration cycles of several seconds. All necessary gating circuits to bring all digital data together into one register are provided. Parity is generated from serial folded code and from parallel-to-serial converted digital words.

All system control circuits are located in the accumulator. These provide the sequential calibration signals and control the time-history generator. The switches are located within the programmer module so that all access to the system in preflight need be made only at the programmer and multiplexer units.

The digital recording is accomplished by gating information from the parallel code register into the record amplifiers which drive a center-tapped special 16-track digital head in a nonreturn-to-zero pattern. The recording of a binary "one" is accomplished by a reversal of current flow in the head to give a change of flux linking the tape. No current reversal is made at the word position if a "zero" is to be recorded. The record amplifier is basically a current driving flip-flop configuration driven in a complementing mode by the code gating pulse.

Analog data recording is accomplished in two possible modes. High-frequency data may be recorded on up to thirteen direct record tracks, and low-frequency signals may be recorded on up to thirteen tracks of FM recording. System data such as servo speed control, wow and flutter reference, pilot's voice, and calibration tones are recorded on one direct record track. Tones for zero and full-scale calibration indications are mixed with the other reference signals during the calibration cycle, which is determined by the accumulator. The tape format is shown in Fig. 4. Fig. 5 shows the analog and digital head layout and track numbering.

The concept of packaging used in this system is Radiation, Inc., standard airborne packaging with minor modifications.[3] The packaging utilizes a card size of four and one half inches wide by five and one half inches deep. The tray that holds the cards varies in length for the various modules depending on the number of cards required. Fig. 6 is a standard module of this system. It can be seen that all cards, connectors, and controls are accessible from one face of the module. The connectors are located on the recessed ledge. The front of the box can be removed allowing access to the cards, test points, adjustments, and local controls (if any).

Fig. 7 is a picture of a card. The cards are clamped with a mechanism which also bows them slightly for use in a vibrational environment. Maintenance can be performed on the cards by the use of extension racks which allow a card to be brought forward out of the box, yet still plug into the chassis wiring. Test points are available on the cards which facilitate the determination of faulty cards in maintenance.

Accessory plugs are provided in each module of the system for attachment of checkout cart electronics. The analog subsystem and the Ampex airborne AR-200 transport are packaged in Ampex standard airborne modules.

DETAIL DESCRIPTION

The most significant points of interest from the standpoint of unique circuit design and logic arrangement are discussed in detail in the following sections.

Low-Level Switch

The low-level switch or gate used is a development outgrowth of the original germanium transistor switch reported by Dorsett and Searcy.[4] Since this time, the improvements made have been those of drive, circuit configuration, and the use of now available silicon transistors. Fig. 8 is a general schematic diagram of the low-level gate. This gate, without any changes, is also used as the high-level gate providing an increase in our system flexibility. Both sides of the double-ended input

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[3] W. A. Corry, "Printed Circuit Packaging for Environmental Extremes," presented at the 1958 National Telemetering Conference, Baltimore, Md.

[4] E. A. Dorsett and J. H. Searcy, "Low-level electronic switch," 1956 IRE NATIONAL CONVENTION RECORD, pt. 5, pp. 57–61.

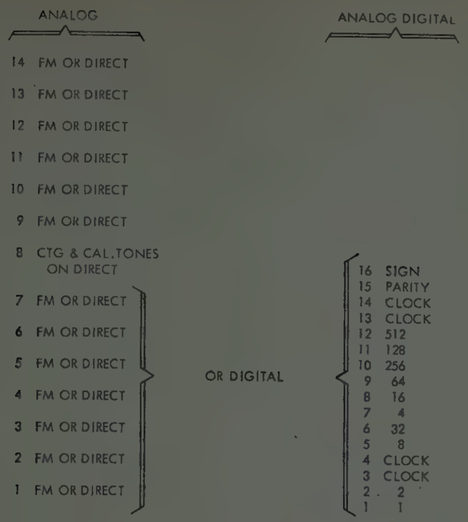
ANALOG
ANALOG DIGITAL
14 FM OR DIRECT
13 FM OR DIRECT
12 FM OR DIRECT
11 FM OR DIRECT
10 FM OR DIRECT
9 FM OR DIRECT
8 CTG & CAL.TONES ON DIRECT
7 FM OR DIRECT
6 FM OR DIRECT
5 FM OR DIRECT
4 FM OR DIRECT
3 FM OR DIRECT
2 FM OR DIRECT
1 FM OR DIRECT
OR DIGITAL
16 SIGN
15 PARITY
14 CLOCK
13 CLOCK
12 512
11 128
10 256
9 64
8 16
7 4
6 32
5 8
4 CLOCK
3 CLOCK
2 2
1 1

Fig. 4—Tape format chart.

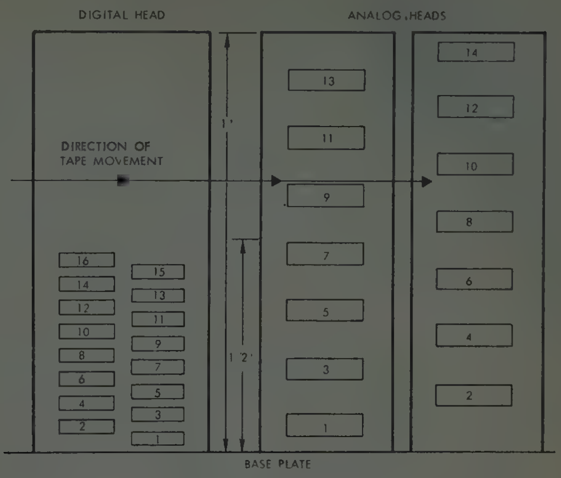
DIGITAL HEAD
ANALOG HEADS
DIRECTION OF TAPE MOVEMENT
1"
1 2"
BASE PLATE
16
14
12
10
8
6
4
2
15
13
11
9
7
5
3
1
13
11
9
7
5
3
1
14
12
10
8
6
4
2

Fig. 5—Head format.

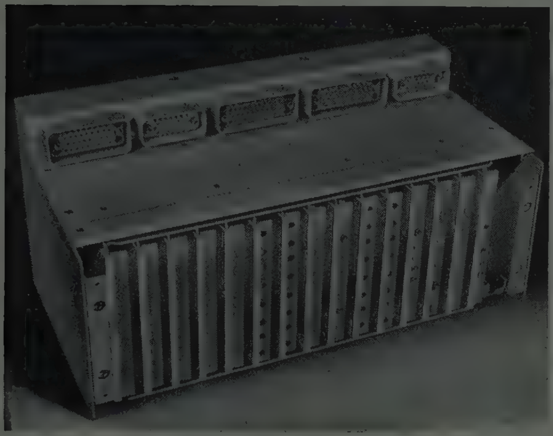

Fig. 6—Multiplexer module.

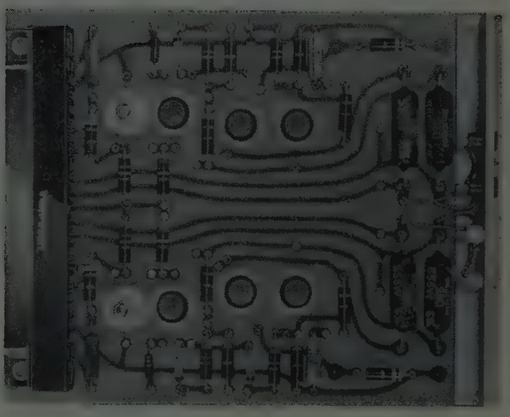

Fig. 7—Photograph of typical card.

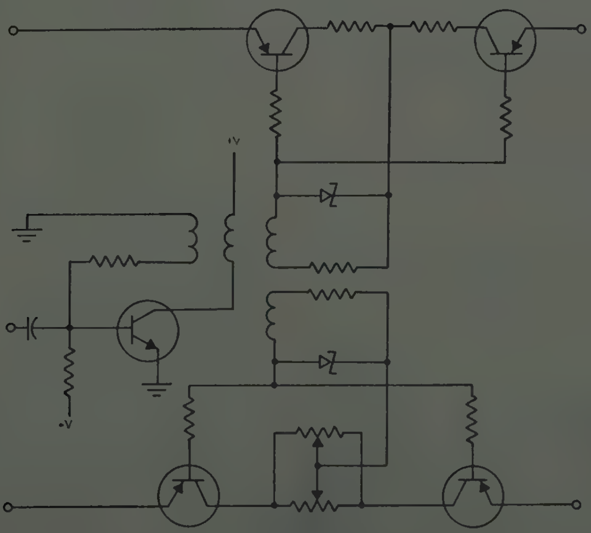

Fig. 8—Low-level gate schematic.

are gated. This is done so that transformers are not required for coupling into the succeeding circuitry. Common mode problems necessitate the use of either a transformer or switching both sides of the line. Transformers which are satisfactory for coupling and yet meet the airborne environmental requirements are not available. The driving circuit is a regenerative transformer coupled amplifier; that is, a blocking oscillator which is turned off before the transformer saturates. Care is taken to maintain the keying signal as a flat-top pulse. This is done by zener regulator circuits in the switch driving circuitry.

There are at least five transistors on the market presently that can be used as the switching transistors. The important parameters are low saturation resistance (low enough so that it is not in an appreciable fraction of the transducer impedance), low-off leakage current (from collector to base) and reasonably low emitter resistance. It has been found that the particular transistors used perform better as a low-level switch if they are used in the inverted connection; that is, the driving signal is applied across the collector-base diode. Each side of the switch contains two transistors. These are provided so that keying signal leakage will be effectively cancelled out. Further, the arrangement of the two transistors "back to back" in each signal line insures that a nonconducting gate will not conduct through forward biased transistor diodes under certain possible cases of voltages applied across the gate terminals.

Even with the availability of silicon transistors some temperature control is still required. However, temperature requirements are not as exacting as they were with the germanium switch. Temperature stabilization is required to minimize keying signal offset compensation drift. This drift has been found to be 0.5 μv per degree Centigrade. It is evident that only temperature excursions of more than 20°C would affect the system accuracy to the extent of one increment or one bit. Controlled heaters regulate the temperature to ± 5°C. Temperature regulation would not be needed in a controlled environment, or in systems which have a resolution of 20 or 30 μv per bit.

Programming

The multiplexer is the programmed type. Programmable means that the sampling rate per channel and the number of channels can be changed and fitted to match a particular application. This type multiplexer permits an acquisition system to be highly flexible. However, when one begins to think about flexibility and its desirability, the problems and consequences of flexibility must also be considered. Flexibility, size, and simplicity of operation often are not compatible. When flexibility and automatic operation are required, size often becomes prohibitive and definitely inconsistent with vehicular system requirements. Therefore, a system that is optimum, that is, a system that meets all requirements in the best possible manner, will result only if the amount of flexibility needed is determined and specified in advance. Often, unwieldy and unusable systems are the result of buyer's specifications being too stringent and the vendor's reluctance or inability to negotiate a suitable system compromise.

The customer for this system, the General Electric Flight Test Facility at Edwards Air Force Base, has been quite cooperative in setting up system specifications and system requirements so that a system could be designed that meets operational requirements and yet is a reasonable size. For instance, flexibility in programming without any restrictions would require that sampling rates of any number be possible. It is immediately obvious that such a system would not be practical. General Electric supplied the required sampling rates. The total number of programs required to be capable of immediate use, and the number of channels and each channel's sampling rate within each of these programs was specified. The sampling rates required in the various programs are always included in the five sampling rates that were originally specified. The most extensive program that might be required at some future time was determined so that the proper amount of hardware could be put in the system. The system requires the capability for a large number of different programs, the ability to change programs in a reasonably short time (five minutes) and that a substantial number of programs be purchased with the original system. To meet the requirements stated above, a program system in which electronics may be moved from logical position to logical position and may be used in a different position for each program was developed. The actual programming is accomplished by changing the chassis wiring of the programming section of the system permitting the system to be relatively inexpensive since the same electronics is used on each program.

Fig. 9 is a block diagram of the programming method used. A countdown from the total multiplexer system sampling rate to the lowest channel sampling rate is provided. In the process of this countdown, care is taken that each of the desired sampling rates of the total system is available. Matrices are attached to the countdown circuit. These provide the output "phases" of the sampling rates desired. The number of phases available at each of these rates is the amount of countdown from the preceding sampling rate. The above described circuitry is common to all programs for the programmer.

The circuitry now to be described changes for each program. Fig. 9 is a block diagram of program X. Fig. 10 is the block diagram of program Y. The number of phases from two adjacent sampling rates are combined in an array of "and" gates so that all possible combinations are made. This produces a number of pulses at the slowest of two sampling rates, and the number of these phases will be the product of the number of the two input sampling rate phases. This mechanism is repeated from the two fastest sampling rates on to the slowest sampling rates. Wherever a particular sampling rate,

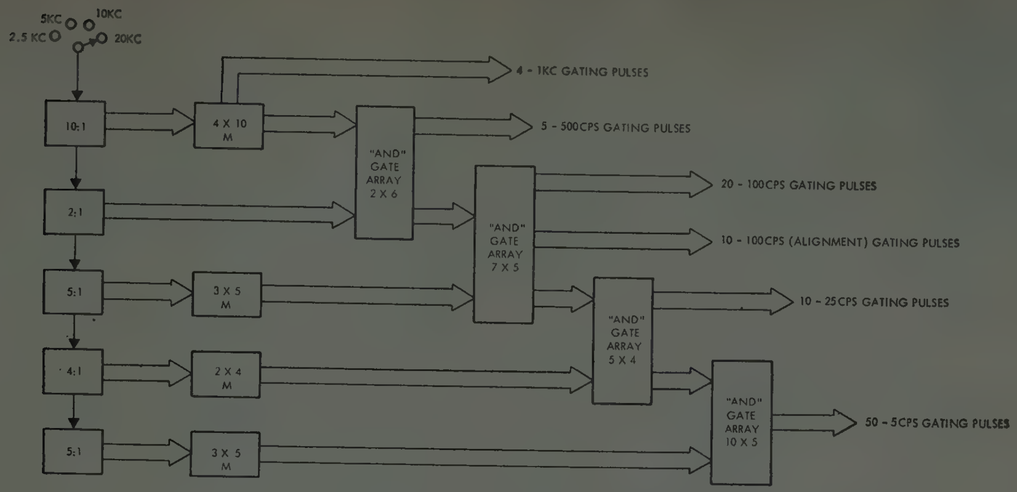
5KC
10KC
2.5 KC
20KC
10:1
2:1
5:1
4:1
5:1
4 X 10 M
3 X 5 M
2 X 4 M
3 X 5 M
"AND" GATE ARRAY 2 X 6
"AND" GATE ARRAY 7 X 5
"AND" GATE ARRAY 5 X 4
"AND" GATE ARRAY 10 X 5
4 - 1KC GATING PULSES
5 - 500CPS GATING PULSES
20 - 100CPS GATING PULSES
10 - 100CPS (ALIGNMENT) GATING PULSES
10 - 25CPS GATING PULSES
50 - 5CPS GATING PULSES

Fig. 9—Programmer with program *X*. Block diagram.

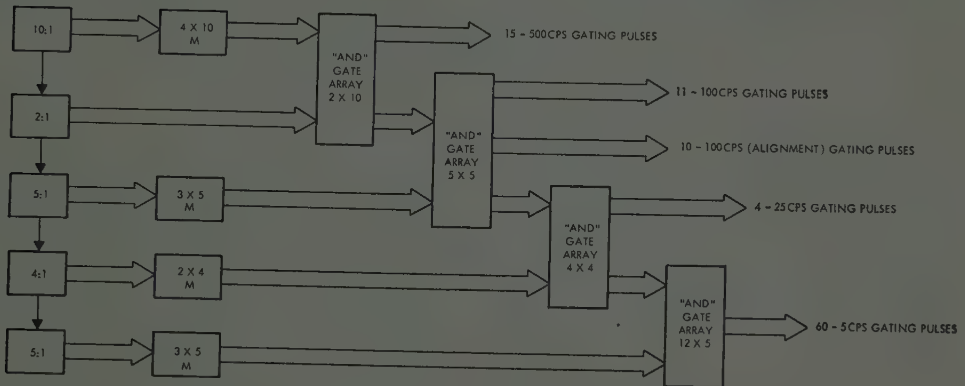
10:1
2:1
5:1
4:1
5:1
4 X 10 M
3 X 5 M
2 X 4 M
3 X 5 M
"AND" GATE ARRAY 2 X 10
"AND" GATE ARRAY 5 X 5
"AND" GATE ARRAY 4 X 4
"AND" GATE ARRAY 12 X 5
15 - 500CPS GATING PULSES
11 - 100CPS GATING PULSES
10 - 100CPS (ALIGNMENT) GATING PULSES
4 - 25CPS GATING PULSES
60 - 5CPS GATING PULSES

Fig. 10—Programmer with program *Y*. Block diagram.

especially one of the higher sampling rates, is desired, it is taken out of the system when generated instead of continuing down the pyramid of "and" gates. Comparison of the block diagrams of programs *X* and *Y* indicates that only rewiring is necessary when changing from one program to another. Rewiring has been accomplished by providing a wired bracket into which the standard cards of the system are plugged. This bracket is separate from the ordinary chassis wiring of the module and it in turn plugs into the module wiring. Therefore, in essence only a portion of the chassis wiring is altered in changing from one program to another. Output of the program wiring to the main chassis wiring determined the position of each input gate or channel which is controlled by a particular program pulse. As set up, this is not to be changed. However, if necessary, the large connector between the program wiring and the main chassis wiring could be altered to change the channel destination of the programmed sampling rate. In fact, the choosing of any ten inputs to be digital instead of analog is accomplished by the rewiring of this connector. The rewiring is facilitated by providing a connector with taper pin connections.

Low-Level Amplifier

A common amplifier is used for fifteen multiplexed channels. Amplification could be done prior to multiplexing through the use of conventional dc amplifiers on each channel. It is apparent that a prohibitively large number of amplifiers would be required. There are basic differences between the signal to the conventional dc amplifier and the signal to the one required in this system. The frequency spectrum of the multiplexed signal is much higher than that of the transducer signal, the source impedance varies from one channel to the next, and the common-mode signal level may vary from channel to channel.

The amplifier shown in Fig. 11 is essentially a direct-connected transistor differential amplifier utilizing periodic clamps for restoration of direct voltage and current levels. The six amplifiers are time-shared to provide the necessary periodic intervals for clamping between sampled data inputs. The input stage consists of a pair of base-fed transistors connected as a differential amplifier. High input impedance and gain stability are improved by inserting resistance in the emitters of the differential pair, while common-mode rejection is

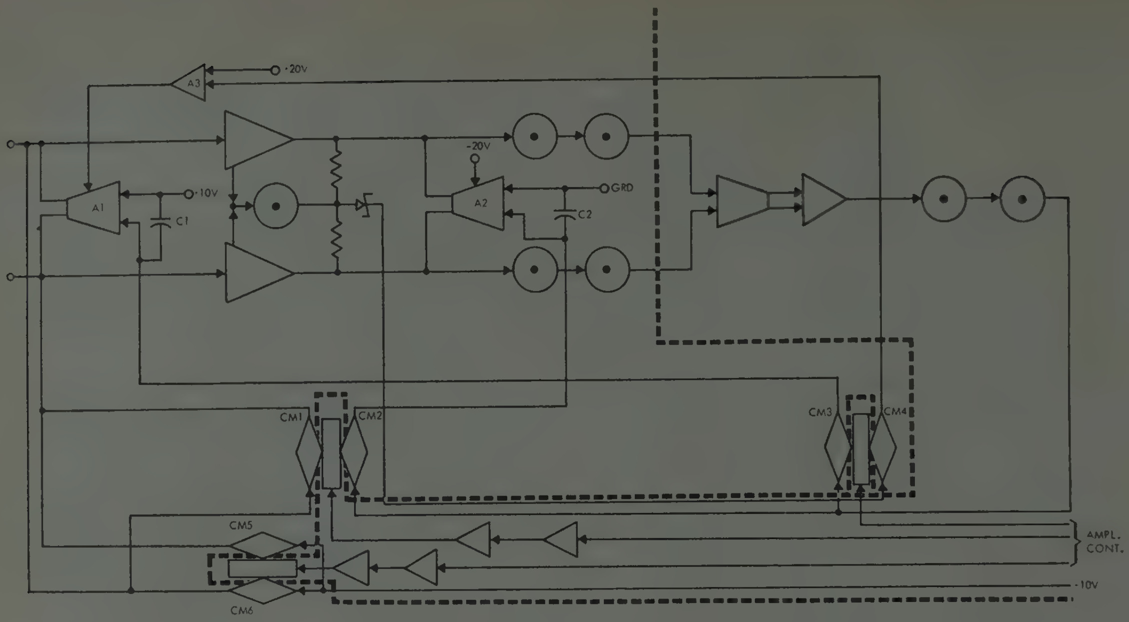

Fig. 11—Clamped amplifier. Block diagram.

gained by feeding the common-mode signal from the common emitter connection to the out-of-phase collector. The input impedance is 30K ohms when the amplifier gain is 500. Noise referred to the input is ten μv and the bandwidth is 100 kc. The input stage is temperature regulated within ±5°C over the ambient range for stability.

The balance of the circuit is basically a group of double-ended amplifier stages with emitter followers for impedance matching forming a feedback amplifier. Drift is prevented by the use of voltage and current clamps in the direct connected circuit. Gate circuits are used to provide the necessary switching functions during the clamping periods. With reference to Fig. 11, the voltage clamp is accomplished by closing two gates, one to short circuit the input (CM1), and the other to connect the output voltage to a comparison circuit (CM2). If the output is not zero, a negative feedback is applied to the collectors of the first stage which drives the output to zero. The voltage at the comparison circuit is stored on C2 so that the level is maintained constant between clamping intervals.

Current drift, as well as voltage drift, may occur in a direct-connected amplifier. With the multiplexer gates nonconducting, the output voltage is switched (CM3) into a differential drive circuit and a constant current is applied at both input circuits and the value of the current to each input is adjusted automatically (A1) to cause the output to go to zero. The total value of the current is adjusted by comparison (CM4) of the first stage output with a known reference in A3. The value of the current is maintained between clamping intervals by storage of the output level on C1.

In addition to the voltage and current clamps, a third restoration process is required. Since the source varies as the channels are sampled, common-mode signals will change with channels. Common-mode crosstalk results due to the requirement for charging distributed capacitance in the gating circuilts. The process consists of bringing the multiplexer output terminals to a voltage more positive (+10 v) than the peak of any common-mode signal with CM5 and CM6, thereby charging the distributed capacity of the gates through the emitter-to-base junction, then returning the clamp voltage to ground prior to sampling.

Four distinct amplifier clamping periods are required: 1) current clamp, 2) voltage clamp, 3) and 4) common-mode clamp. Programming is arranged so that the four clamp periods are available in two sampling periods between data samples while a different amplifier is activated.

Analog-to-Digital Converter

The analog-to-digital converter (coder) used in this airborne acquisition system is the Radiation Inc. standard product designed for use in vehicular environments. This airborne coder is fully transistorized, is ruggedized sufficiently well to withstand sled environments as well as airborne environments, and is composed of silicon transistors throughout. Earlier models of the coder have been discussed in several papers[5,6] in the literature previously. Its mode of conversion is that which is popu-

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[5] L. S. McMillian, "Development of a high-speed transistorized 10-bit coder," 1957 WESCON CONVENTION RECORD, pt. 5, pp. 73–76.

[6] R. E. Marquand and W. T. Eddins, "A transistorized PCM telemeter for extended environments," 1957 WESCON CONVENTION RECORD, pt. 5, pp. 76–81.

larly known as "Half-Split Technique" or as a digital-to-analog converter with digital feedback.

The coder contains the conventional Radiation Inc. sample and hold circuitry modernized to meet the new vehicular requirements. Earlier models or configurations of this sample and hold circuitry have been discussed in the literature.[7] The ouput of the coder is in the form of a serial code with available word and frame sync, since the coder is a basic unit performing enough functions for use with an RF link. Provision has been made to limit the full scale of the coder to less (±1008) than that possible with eleven bits (±1023). Other functions which might be considered to be part of the coder, such as the generation of parallel code, parity, and folded code (10 bit plus sign) are performed in the accumulator.

Fig. 12 is a block diagram of the coder. It can be operated in either the free-running (which means continuous) or commanded mode. The rate at which the binary digits are coded is kept constant at 280 kc. It takes 43 μsec to code each word. Since this system operates at 20 kc, 10 kc, 5 kc, or 2.5 kc, the coder is commanded every 50, 100, 200, and 400 μsec respectively. The coding time is fixed at 43 μsec and the remaining μsec before the next coding period is dead time.

The logical decisions to retain or reject the binary weighted reference currents while sending in the next smaller current are made in sequence. Sequential pulses on separate lines are therefore necessary. A feedback Gray code counter is used for this application rather than the ordinary binary counter. This type of counter has the following advantages: 1) there is no delay of forward carry as in the binary counter, and 2) the fastest operating flip-flop need run at a rate of only one half that of the corresponding flip-flop in a binary counter. The electronics necessary for feedback is available with either counter since a sorting matrix is necessary to obtain the number of pulse lines desired. A broken ring counter was not used because of the amount of electronics required.

A method of describing the conversion cycle is to state that the coder generates a reference analog within itself in successively smaller binary steps, these reference analogs being summed together with the input analog current-wise and then successively compared to signal ground. The logical operations involved in the coder are to generate a binary weighted reference analog, sum it with the input analog, compare it to signal ground, reject or retain this first reference analog dependent upon the comparator's decision, and switch into the reference analog just determined, another binary reference analog of half the magnitude of the preceding reference analog. Thus, the coder successively tried to match (in opposite polarity) the input analog in binary steps from most significant to least significant by retaining some steps and rejecting others. Fig. 13 is a schematic of the summing network. It is a current device and as such contains precision resistors so that accurate weighted currents may be summed. The voltage from which these currents are generated must be of a greater precision and accuracy than the least significant bit of the number of bits generated in the coder. Transistors are used to switch the precision summing currents into the current summing network.

An analog-to-digital converter of the type just described (or most types) must contain analog circuitry that is more accurate than the number of digits or binary increments into which the input analog can be quantized. Therefore, an eleven-bit coder or ten-bit plus sign, where the eleven bits are binary, implies 2047 possible quanta against which the input analog is compared. One part in 2000 is equal to 0.05 per cent of full-scale accuracy. Therefore, the analog circuitry such as the reference power supplies and sample hold circuitry must be accurate to some greater precision. The analog portions of this coder meet these requirements.

Since the coder must maintain extremely accurate analog signals, some temperature control is required. Such circuits as the comparator, sample and hold, precision reference power supplies, and the most significant digits of the summing network must be temperature controlled to some extent. The control required to maintain sufficient accuracy is plus or minus five degrees Centigrade over the range of ambient temperature excursion. Heaters and control circuitry are provided in the coder.

Accumulator

A block diagram of the accumulator is shown in Fig. 14. The accumulator unit performs the following functions:

1) Folding code
2) Serial-to-parallel conversion
3) Parity generation
4) Parallel-to-serial conversion
5) Unique word generation
6) Word rate and time history
7) Digital input gating
8) Gray-to-binary conversion
9) Parallel storage
10) Recording
11) Calibration control

Most of the circuitry involved is conventional digital type configurations, but it is felt that the arrangement of these circuits to perform the listed functions represents a maximum usage of a minimum number of components.

The logic for folding (transforming an eleven-bit code to ten-bit plus sign) the serial code from the coder is an inverse exclusive "or" circuit. The simplified Boolean expression to be mechanized is $AB+\overline{(A+B)}$ where A

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[7] W. T. Eddins, "Sample-and-hold circuits for time correlations of analog-voltage information," 1958 WESCON CONVENTION RECORD, pt. 5, pp. 21–26.

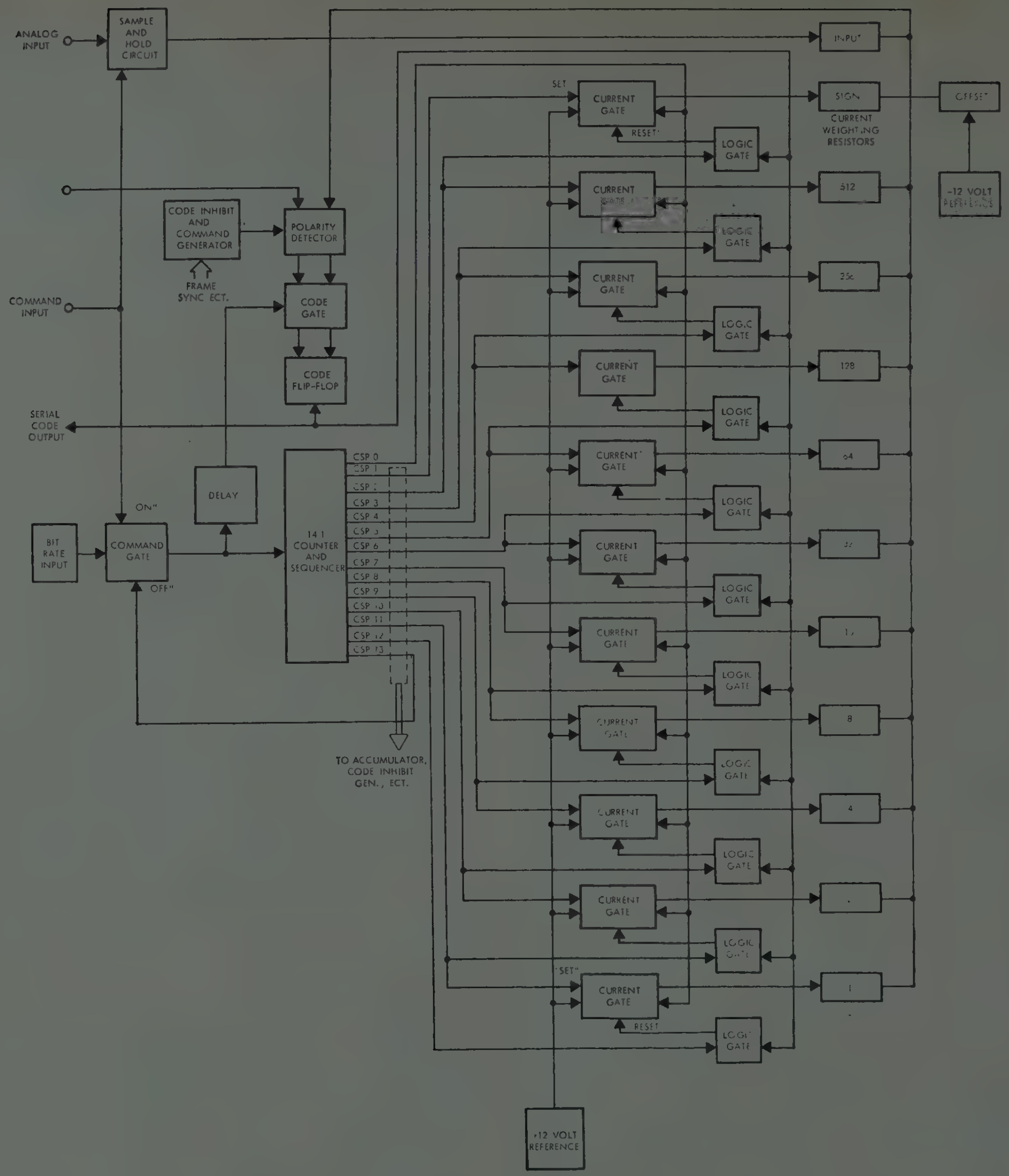
ANALOG INPUT
SAMPLE AND HOLD CIRCUIT
CODE INHIBIT AND COMMAND GENERATOR
FRAME SYNC ECT.
POLARITY DETECTOR
COMMAND INPUT
CODE GATE
CODE FLIP-FLOP
SERIAL CODE OUTPUT
DELAY
"ON"
BIT RATE INPUT
COMMAND GATE
"OFF"
14:1 COUNTER AND SEQUENCER
CSP 0
CSP 1
CSP 2
CSP 3
CSP 4
CSP 5
CSP 6
CSP 7
CSP 8
CSP 9
CSP 10
CSP 11
CSP 12
CSP 13
TO ACCUMULATOR, CODE INHIBIT GEN., ECT.
SET
CURRENT GATE
RESET
LOGIC GATE
INPUT
SIGN
CURRENT WEIGHTING RESISTORS
512
256
128
64
32
16
8
4
2
1
OFFSET
-12 VOLT REFERENCE
+12 VOLT REFERENCE
"SET"

Fig. 12—Coder. Block diagram.

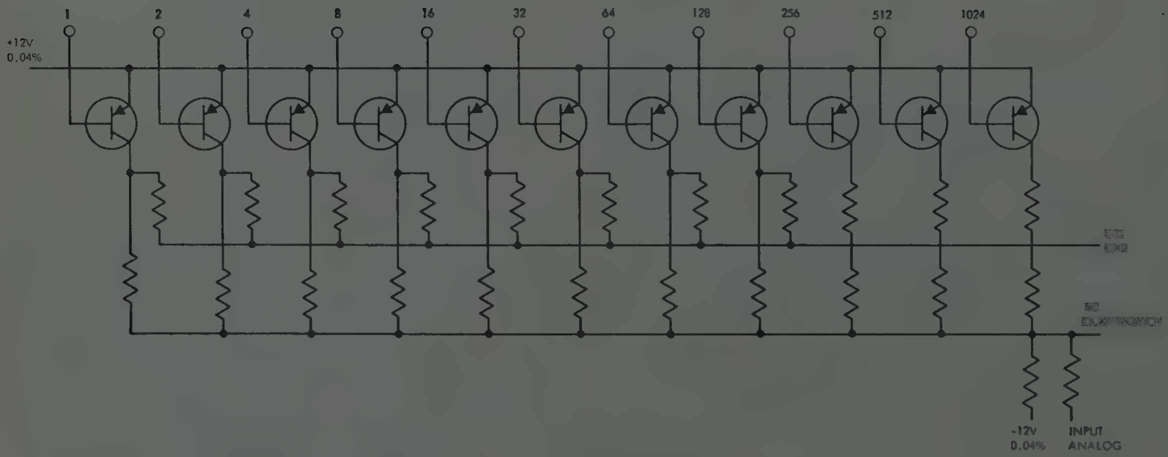
1
2
4
8
16
32
64
128
256
512
1024
+12V 0.04%
SIG GND
TO COMPARATOR
-12V 0.04%
INPUT ANALOG

Fig. 13—Current summing network. Schematic.

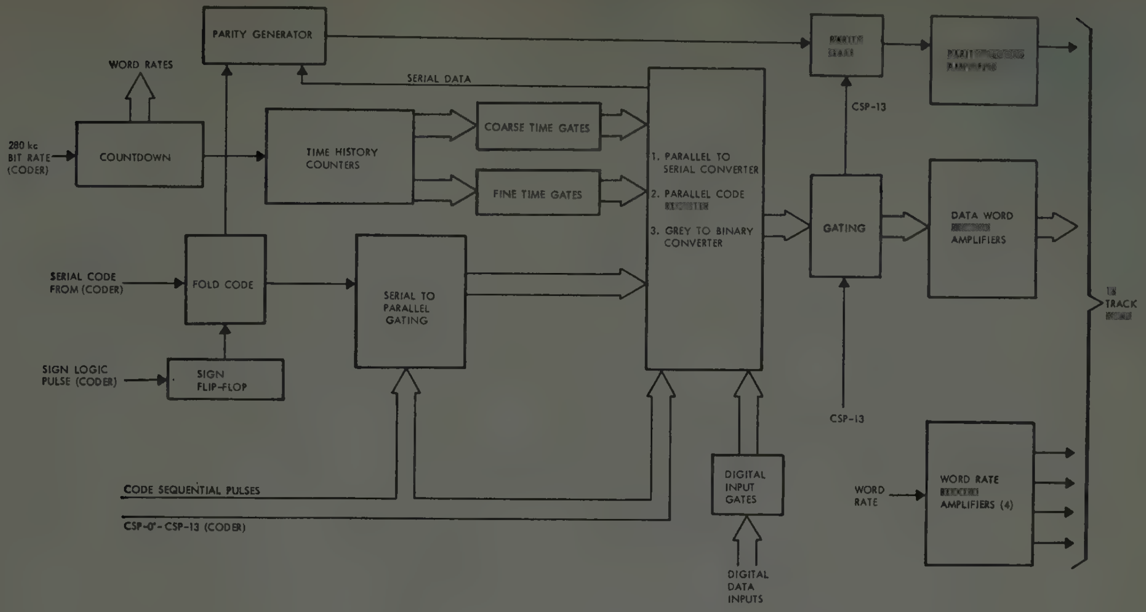
PARITY GENERATOR
WORD RATES
SERIAL DATA
280 kc BIT RATE (CODER)
COUNTDOWN
TIME HISTORY COUNTERS
COARSE TIME GATES
FINE TIME GATES
1. PARALLEL TO SERIAL CONVERTER
2. PARALLEL CODE [redacted]
3. GREY TO BINARY CONVERTER
[redacted]
[redacted]
CSP-13
GATING
DATA WORD [redacted] AMPLIFIERS
SERIAL CODE FROM (CODER)
FOLD CODE
SERIAL TO PARALLEL GATING
TRACK
SIGN LOGIC PULSE (CODER)
SIGN FLIP-FLOP
CSP-13
CODE SEQUENTIAL PULSES
CSP-0"-CSP-13 (CODER)
DIGITAL INPUT GATES
WORD RATE
WORD RATE [redacted] AMPLIFIERS (4)
DIGITAL DATA INPUTS

Fig. 14—Accumulator. Block diagram.

and B are the sign and data word, respectively. A minimum of logic elements is used as shown in Fig. 15. The purpose of folding code is to give a representation of numbers which is easily understandable by the human operator. The code zero corresponds to zero signal and the code value increases from that point with any change in signal and the most significant bit indicates the sign.

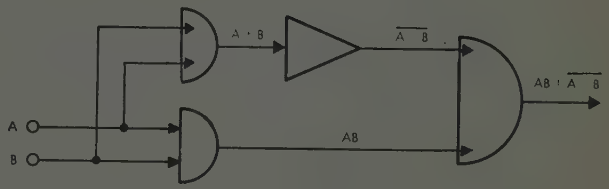
A · B
$\overline{A\ B}$
$AB + \overline{A\ B}$
A
B
AB

Fig. 15—Fold code logic. Block diagram.

Serial-to-parallel conversion is accomplished by "and" gating the sequential pulses generated in the analog-to-digital converter with the folded serial code into the parallel storage register.

Parity generation for the serial code word is accomplished by gating the delayed clock from the coder with the code to drive the complement input of the parity flip-flop (PFF). Since the PFF is preset to a fixed condition prior to the occurrence of each data word, sampling the condition of the PFF at the end of the word allows the parity bit to be set to a "one" if the "ones" total is even. Parity computation for the digital inputs which are introduced into the parallel code register broadside is accomplished with a parallel-to-serial conversion and "one" counting in the same circuit. The coder output is inhibited during this process.

Unique words are used as markers in the system to indicate to the ground data processing that some event has taken place. A specific word, above the limited full-scale code, is recorded on the tape in a unique time slot to indicate: 1) beginning of a frame of data, 2) beginning of a data cycle, 3) calibration (zero and full scale), and 4) other events as required by system operation. The markers are generated by gating a voltage (indicating a "one") into the proper bit of the parallel code register with the proper program timing pulse.

Digital counting techniques are used to achieve the various word rates and the time-history rate of 100 pps. Synchronous, stable pulse frequencies are obtained from the single crystal-controlled oscillator in the coder. The same source is used for the coder bit rate. A selector switch at the programmer module is used for selection of the various word rates which drive the programmer counters and control the coding process. The time-history is an accumulation of binary counts in two 1000 to 1 counters. The fine and coarse time words represent a maximum accumulation of ten seconds and 10,000 seconds, respectively. The two time words are gated into the parallel storage register with the proper timing pulses from the programmer.

Gating of the inputs from digital shaft encoders is accomplished simply by "or" gating the ten-bit parallel signals into the parallel storage register. The timing is controlled from the programmer digital query driver, and any channel time position may be used for digital information because of the ability to "patch" timing pulses to the encoder drivers.

Since the most desirable digital shaft encoders generate the digital data in Gray code, it becomes necessary to convert to binary coding prior to recording. A very economical solution was found since sequential coding pulses were available in the form of the CSP's from the coder. Fig. 16 shows the logic arrangement for the conversion process. It is significant to note that no intermediate buffer is required, resulting in minimum circuitry.

The remaining function of this component of the system is recording, which has already been described in detail.

SYSTEM OPERATION

Operational flexibility is greatly enhanced by the modular construction of the components of the system. Flight testing of the smallest aircraft or larger missiles to the largest aircraft may be accomplished with ease simply by using those portions of the system required; that is, a smaller number of multiplexer modules. The flexibility of programming allows an operational versatility never before offered in airborne PCM systems.

If telemetering provisions are required in addition to or in place of tape recording, an FM transmitter may be connected to the serial output of the coder. A unique word for frame synchronization and alternating bit preceding each word for word synchronization can be provided. The resultant telemetric data conform to the standards set forth by the telemetry working group of the IRIG.

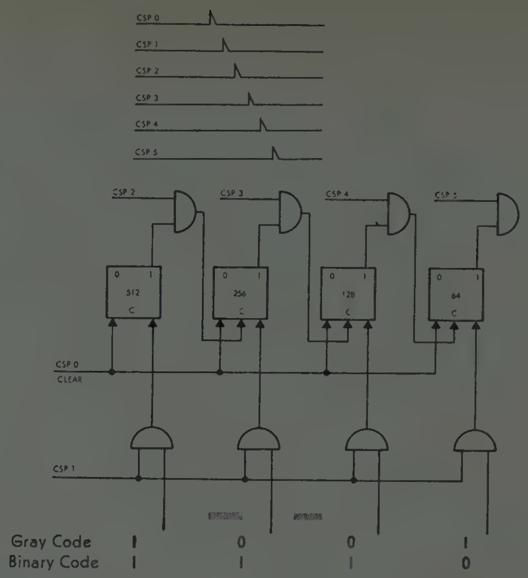

Fig. 16—Gray-to-binary converter. Block diagram.

ACKNOWLEDGMENT

Efforts were required of many of our associates in performance of system and component design. L. S. McMillian, W. T. Eddins, W. A. Corry, G. E. Bradford, and R. E. Marquand are a few who have made significant contributions to the success of the project.

Transistorized Motor Speed Controls for Satellite Tape Recorders*

MERWIN B. PICKOVER†

Summary—**The proposed paper discusses the problems arising in the design of an ac transistorized power supply for low power synchronous motors used in satellite recorders. It treats a current design for a specific satellite application, presents the design of the over-all system, and states alternate methods that can be used. The paper also discusses the over-all system parameters which must be considered in the choice of a hysteresis synchronous motor, the motor's operating frequency, and its characteristics.**

Also treated in the paper is a transistorized dc motor control used in a Signal Corps tape recorder. It discusses the servo loop used to accomplish the control of speed, and the accuracy that may be obtained. Diagrams of the exact circuit are presented and analyzed so that the reader may apply the technique to his future needs.

The relative advantages and disadvantages of the dc motor and ac synchronous motor in the low-power output range are discussed.

The system is designed to meet the environment of the satellite package. The emphasis is on a system requiring a minimum of power consumption and smallest possible size and weight, with no sacrifice of reliability or performance.

INTRODUCTION

SINCE size, weight, efficiency, and reliability are of primary importance in satellite circuitry, the government and private industry have been seeking a small transistorized speed control in order to maintain the speed of small dc and ac motors used for tape recorders and other motor applications.

DISCUSSION

At the present time the use of ac synchronous motors for extremely low power level may become impractical, due to the fact that a stable ac power supply and amplifiers are necessary for primary power, and the over-all efficiency of the system therefore suffers.

The inherent starting torque of the small dc motor is greater than an equivalent output hysteresis synchronous motor. This and its greater efficiency, are advantages that must be considered.

The use of mechanical speed governors to maintain a constant speed in the dc motors has been ruled out for many applications, due to the inherent vibration problems. The constant making and breaking of the governor contacts cause noise effects which must be suppressed.

At present four methods of control for small dc motors have been exploited. One is to place a dc generator at the end of the shaft whose output dc is now proportional to speed. By comparing this output to a reference (Zener diode), one can design circuitry to obtain the desired speed control.

An alternate method is to place another winding in the slots of the armature and bring out this ac voltage to slip rings. The frequency of the output voltage $f=np/120$ (n = speed in rpm and p = number of poles). Since this method requires another set of brushes, it is undesirable.

Still another method utilizes commutation ripple frequency which is proportional to speed. This method is not too successful at present, but it certainly should be explored.

The fourth method and the one most widely practiced is to place an ac generator at the end of the motor shaft whose output frequency is proportional to speed. A discriminator converts this change in frequency to a change in dc voltage, which is then applied to a dc amplifier to the motor input, or field winding.

DC Motor Speed Control

With the above considerations in mind a package of approximately three cubic inches, with a total power consumption of less than 100 mw for control purposes, was designed. Silicon transistors may be used throughout as a precaution for high temperature operation although the range of temperatures encountered is relatively small because of the design of the satellite package. The complete wiring diagram is illustrated in Fig. 1.

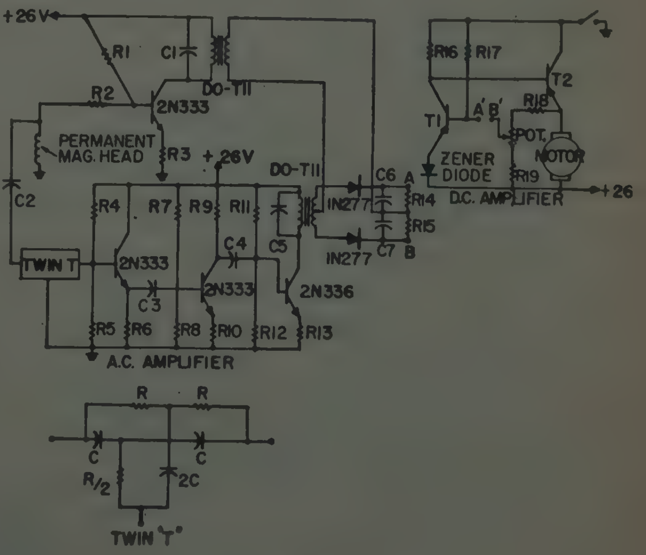

Fig. 1—DC motor supply.

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† U. S. Army Signal Research and Development Lab., Fort Monmouth, N. J.

Principle of Operation: In order to control the speed of the dc motor a block diagram is illustrated in Fig. 2. The ac generator is mechanically driven by the dc motor. The output of the ac generator is fed into a sensing circuit. The construction of the ac generator is described later in the paper. A twin tee is used as the reference for frequency. The twin tee contains vitamin *Q* capacitors and one per cent resistors. These components were then selected so that the twin tee characteristics produced a minimum null at the desired frequency and a minimum drift of the null with temperature variations. Its characteristics are as shown in Fig. 3. An emitter follower was then placed across the output of the twin tee, in order to transform the relatively high impedance of the twin tee to a lower value. The twin tee output is then amplified by two stages of RC coupled amplifiers to give a 90° phase shift at the desired frequency and with additional gain. The output of the RC networks is now compared to the generator output, as illustrated by Fig. 4. This discriminator whose output has the sensitivity of one volt per 3 cps is linear to about ±2 volts dc.

The dc amplifier maintains the nominal dc voltage necessary for operation. The discriminator output adds to or substracts from this nominal.

AC Generator: A problem that arose was the inability of permanent magnetic designers to produce a small magnet with the desired number of poles which would produce a relatively high frequency at a relatively low speed. The conclusion reached is that it is impossible to construct a magnetizer for this application.

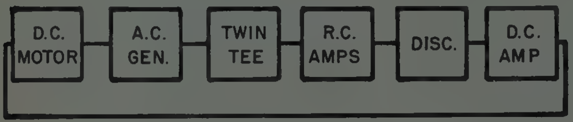

Fig. 2.

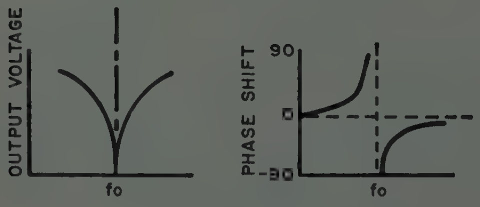

Fig. 3.

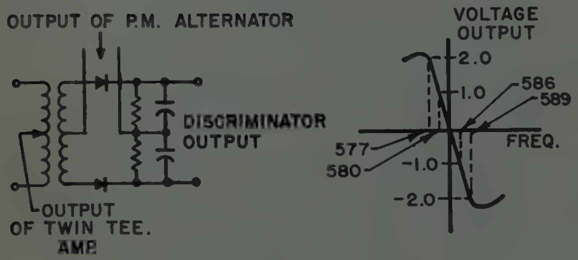

Fig. 4.

In order to overcome this a generator was constructed. This generator consists of a rotating magnesium wheel with permanent magnet inserts and a pick-up coil.

Permanent magnets: cunife;
Size of magnets: $\frac{1}{8}$ inch long, $\frac{1}{16}$ inch diameter;
Number of magnets or poles: 14;
Speed of dc motor or ac generator: 5000 rpm; frequency of output voltage $= np/120 = 583$ cps;
Pick-up coil of the generator: 12,000 T #43 HF wire;
DC resistance: 2000 ohms; 5000 ohms ac impedance;
Generator air gap: 0.010 ± 0.002;
Pick-up core material: silicon steel, U. S. Transformer #52 or equivalent;
Size of wheel: $\frac{7}{8}$ inch O.D.;
Frequency of twin tee: 583 cps;
Determined by: $\omega = 1/RC$, $\omega = 2\pi f$, $R = 13.3$ k, $c = 0.02$ μfd, $f = 583$ cps.

The value of R is usually chosen so that the twin tee does not load down the permanent magnet alternator winding, or an emitter follower may be necessary between the output of the alternator and the twin tee.

DC Amplifier: This circuit can best be explained by a typical example.

Given

1) nominal motor voltage 16 vdc,
2) maximum motor voltage 18 vdc,
3) minimum motor voltage 14 vdc,
5) nominal motor current 60 ma,
5) battery voltage 26 vdc.

Insert the motor in the circuit between the emitter of the *p-n-p* power transistor and +26 volts as shown in Fig. 1.

The nominal voltage at the emitter of the *p-n-p* power transistor is approximately +10 vdc.

The base of the *p-n-p* power transistor is approximately at +9.5 volts. That is also the potential at the collector of $T1$. $T1$ should be biased for 4.0-ma collector current. $R16$ can now be chosen as approximately 3000 ohms for steady-state conditions.

The potentiometer across the motor in conjunction with the Zener diode in the emitter circuit provides the desired dc level so that the output from the discriminator may be applied directly; it also provides a high impedance to the discriminator so that its loading effect may be neglected. A voltage more proportional to any discriminator output voltage, + or −, appears across $R16$ and therefore across the motor. The dc gain for the circuit shown is approximately one.

The Zener diode in the emitter is chosen to maintain a constant 14-volt drop and brings the base of $T1$ to approximately +11.5 volts. The potentiometer across the motor is set so that its arm is at this exact potential.

Therefore, points A' and B' are at the same dc potential and the detector may be impressed across A and B without difficulty.

The circuit is versatile. If the polarity of the 26 volts is reversed, a simple change of equivalent *n-p-n* transistors to replace the *p-n-p* transistors and a reversal of the Zener diode polarity will produce identical operation. The use of *p-n-p* transistors in the ac detector circuit gives similar results, with a change in battery voltage polarity.

Temperature compensation can be accomplished by a NTC for *R*18.

The accuracy of the speed control system is predicted in the following manner. It is necessary to know the change in motor voltage necessary to maintain a constant speed at all loads and ambient variations. By simply comparing this change in voltage to the change in detector dc voltage output, one can see what change in frequency to expect. If the detector output vs frequency is compensated for temperature variations, one can predict the change in speed that will occur. The present system is good for approximately one per cent total speed variation, for all ambients between −10°C to +60°C. It can be improved if the application presents itself.

AC Synchronous Motor Supply

If one is willing to sacrifice efficiency, quick starting, and size, the hysteresis or other synchronous motors have numerous advantages over the dc motor. Among them are longer life, no brush commutation noise and no brush dust problems. In order to produce the ac power, many considerations must be taken into account. The designer must choose a standard for frequency, a method of countdown, and a power amplifier. The choice of an operating frequency is usually governed by the synchronous motor manufacturers. At present 60 cps and 400 cps are commonly produced, although prototypes are available for 125 cps as well as for higher frequencies. The reference may be a crystal or tuning fork, or, if extreme accuracy is not required, a transistor oscillator may be used. When using a crystal, there is the problem of numerous countdown circuits and possibly magnetic memory cores for storage of pulses. When using a low-frequency tuning fork, countdown by four or eight is usually sufficient. In satellite application the over-all-system must be considered. If a tuning fork is used, it can be used for reference for other circuitry and also can be the reference for the synchronous power supply as in the case mentioned. The choice of a synchronous motor frequency presents the following problem. A 400-cps motor will be smaller in size and have a greater efficiency than an equivalent output 60-cps motor, but one must realize that this change in operating speed presents a mechanical problem of gearing or belt reduction in the recorder if low-speed operation is desired. The assumption is made that the number of poles is constant. The problem of the motor designer is to produce a small motor punching with many teeth, in order to keep the speed down to a reasonable operating point for a given frequency.

With these considerations in mind a system was designed to produce 125-cps power for a synchronous hysteresis motor for a tape recorder. (Refer to Fig. 5.) The tuning fork was manufactured by Philmon Laboratory. The synchronous motor was manufactured by Herbert C. Roters Inc., a recognized expert in the field of high efficiency hysteresis synchronous motors. All other circuitry was designed by the U. S. Army Signal Engineering Laboratory. It consists of a tuning fork and oscillator circuitry, two multivibrators used as countdown by four, a driver, and a push-pull amplifier. The output voltage to the synchronous motor was 12 volts rms ±2 per cent for ambients of −10°C to +60°C. Extreme accuracy in output voltage was not a requirement, since the motor will stay in synchronism for voltages between 11 to 15 volts with no degradation in performance. However, the higher voltage produces higher losses in the hysteresis motor, and actually a waste of power.

The multivibrator and push-pull output transistors were matched pairs. Temperature compensation can be accomplished by an NTC for resistor *R*28. The tuning fork oscillator and multivibrators required less than 100 mw of power, and the electrical efficiency of the driver amplifier and push-pull output was approximately 55 per cent. The transformer efficiency was slightly greater than 80 per cent. Class AB push-pull operation and the transformer efficiency were the limiting factors which controlled the over-all system efficiency.

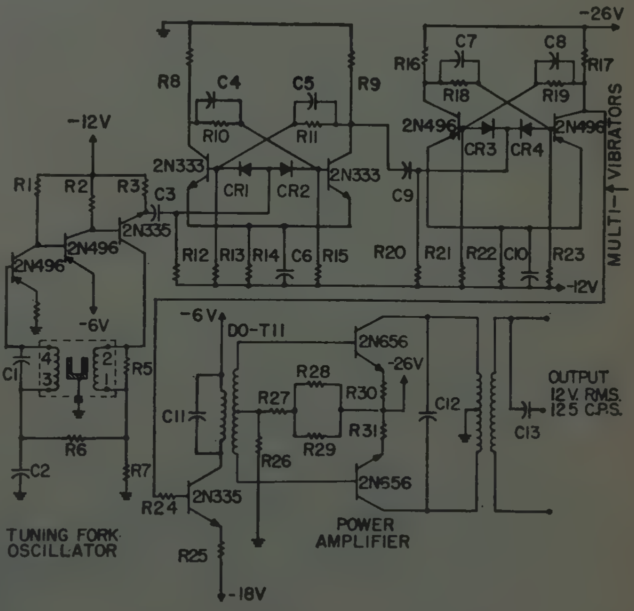

Fig. 5—AC synchronous motor supply.

CONCLUSION

The author has presented two of the circuits used for the control of the speed of small motors. In particular, the controls have been used for a tape recorder, record cycle, and playback cycle. The type of control lends itself to other satellite application, when the control of the speed of motors at all environments is necessary. Most of the components have been selected. All transistor characteristics were observed on a curve-tracer to insure the reliability of the package. As stated above, certain transistors were matched pairs, so that their relative changes in gain vs temperatures would be minimized. If one is striving for simplicity of circuitry, minimum power consumption, minimum size, and reliability, the above procedure must be observed.

ACKNOWLEDGMENT

The author wishes to express his appreciation to Dr. R. A. Hanel, formerly of the U. S. Army Signal Research and Development Laboratory, now affiliated with the National Aeronautics and Space Administration, for his technical contribution during initial design of the system.

Checkout and Countdown of the Larger Space Probe Missiles*

W. O. CAMPBELL†, MEMBER, IRE

Summary—**While it is obvious that missiles for different space programs carry widely different electronic equipment, it is not so apparent that missiles on the same program may carry just as wide a variety of equipment. Checkout and countdown equipment must be versatile, not only because of the variety of electronic airborne equipment, but because of the special needs of research and development programs and for follow-on programs. Moreover, the equipment must offer a high level of performance reliability, so that the operator may concentrate on firing the missile.**

The design of the centrally programmed checkout and countdown set presented here incorporates digital techniques. The punched tape supplies those commands that are subject to frequent change, such as GO—NO-GO limits—and mode and range commands for the parameter converters. Manual operation and visual evaluation are possible. These provisions also raise follow-on utility.

Use of the design criteria of self-verification at every step, positive malfunction localization, standby verification and design simplicity yield a design with a high level of performance reliability.

INTRODUCTION

THE missiles of today, as they probe deeper into space, are forced to carry a greater and greater variety of electronic and research equipment, necessary because we know so little about the phenomena that occur in outer space.

The stages of the missile also undergo changes in the electronic configuration, from flight to flight. Each change, however minor, usually calls for a similar one in the support equipment. A change in autopilot mixing ratios or gain settings requires a corresponding change in the checkout equipment. More often than not, the change is of greater magnitude and expense than the original change in the missile. Even so simple a change as the tolerance of a particular parameter of the missile requires a corresponding change in the GO—NO-GO testing limits.

This presents a never-ending series of new problems to the launch crew. To make the situation more interesting, the crew does not have all the time in the world to make changes. With the next flight date only two weeks away, they can hardly be expected to spend their time reworking the support equipment. Obviously, the design must be such that changes in it will not impede the steady flight-to-flight schedule of testing.

On top of all these requirements, performance reliability is required of the checkout and countdown equipment. These two requirements are of paramount importance: 1) performance reliability and 2) versatility. The following discussion develops these requirements more fully and presents a block diagram solution to them for the checkout and countdown of a typical deep space probe missile.

PERFORMANCE RELIABILITY

The checkout equipment designs of several years ago were predominantly manual in nature, consisting of standard laboratory meters, oscilloscopes and other familiar equipment. Being manual, they were quite reliable. Certain problems were presented, however, usually involving personnel training and speed. Gradually, the equipment evolved into semi-automatic designs. The evolution was encouraged on the part of the military by

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† Electronics Support Equipment, Space Flight Div., The Martin Co., Baltimore, Md.

the allocation of sums of money for ground support research and development.

In converting from the manual to the semi-automatic designs, it was desirable that the high reliability level of the manual equipment be preserved. This is best stated in a rather paradoxical fashion; *i.e.*, the countdown reliability of the support equipment should be greater than that of the missile.

Gradually a more specific term, "performance reliability," evolved. Performance reliability means that, considering the system as a whole, the support equipment must always correctly evaluate the situation for the operator. This characteristic quickly instills a high level of confidence in the equipment.

Fig. 1 explains the characteristic more fully, showing what it is and how it is secured. Follow the path of a typical signal coming from a missile and arriving at the comparator. It goes first through a self-test relay contact, then through a programmer and finally into a signal conditioner, which conditions the signal to a common level so that a single comparator can be used for the majority of the tests.

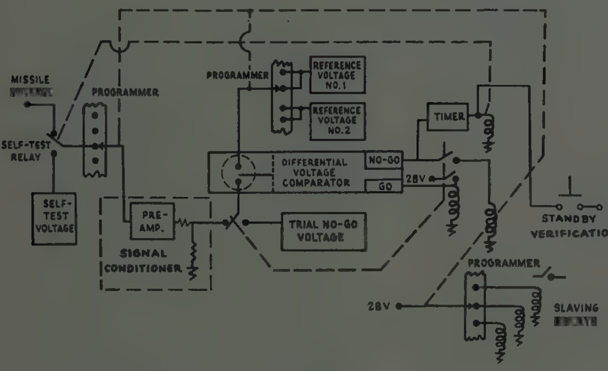

Fig. 1—General method of automatic analog testing.

The programmer, shown as a simple stepping switch, also connects a proper reference into the other side of the comparator. The one shown is a fixed-difference comparator such that, if the difference between the conditioned signal and the reference value is less than a predesigned amount, the comparator will register a GO condition.

Assume, for purposes of this discussion, that this condition obtains and that the comparator is registering a GO condition. If asked to make a guess, the reader would assume that the next operation would be that of the programmer advancing to set up the next step. Such is not the case here. Instead, the equipment temporarily memorizes the apparent fact of the GO condition and sets it aside, because it is not certain at the moment whether the GO condition is bona fide or the result of the combination of a faulty missile and malfunctioning checkout equipment.

The next operation is that of programming a trial NO-GO voltage into the comparator. Now, if the comparator can be made to recognize a GO condition, which has just occurred, and likewise a NO-GO condition, occurring at the moment, then the comparator decision has been verified. The GO condition previously received is a bona fide GO condition and is recognized as such. At this point, a signal is flashed to the programmer to advance to the next step. This process is called self-verification at every step. It represents the first part of certain design criteria which give this method of testing a very high performance reliability.

If, on a particular test, a sustained NO-GO condition is registered by the comparator, a timer is energized (see Fig. 1). After a short interval, the timer activates a self-test relay. The relay temporarily disconnects the missile response and in place of it connects a design-center response. If the checkout equipment is operating properly, it will recognize the design-center value as a GO condition. This response, by means of circuitry not shown, is used to light a lamp or other signalling device to indicate that the trouble lies within the missile, since it is not in the checkout equipment.

If the comparator gives a NO-GO condition upon evaluation of the design-center value, the trouble lies within the checkout equipment. An appropriate signalling device is energized to signify the fact. This process is positive malfunction localization, a part of the normal functioning of the equipment. It is the second of the design criteria which must be used to give performance reliability.

A further requirement is the certain knowledge that the equipment is capable of making a countdown before a scheduled countdown takes place. It is necessary to use the equipment on a standby basis and verify that it is operating properly. This is done in Fig. 1 by closing a standby verification switch to activate all of the self-test relays and by allowing the checkout equipment to evaluate the design-center values throughout. Completion of the sequence verifies that it is operating properly at all steps. Automatic standby verification is the third criterion for performance reliability.

The fourth criterion states that the first three must be incorporated into the basic design with addition of very little extra equipment. Obviously, this is a criterion; to do otherwise would cause the over-all reliability to go down rather than up. Examination of Fig. 1 shows that this last criterion has indeed been met. The functional modules shown are present in one form or another in any semi-automatic checkout design. The difference lies in the logical way the modules are used.

VERSATILITY

Several kinds of versatility are required. One is electronic versatility commensurate with the variety of electrical and electronic gear. The first flight may carry a standard guidance system for coordination purposes. A pre-programmed autopilot may be used as an interim model. On the second flight, the standard system may be used to guide the missile. Certain new communications equipment may be aboard. The same autopilot as before is used, but now the gains, tolerances, and mixing ratios are different. These factors all add up to the requirement that checkout equipment designs incorporate a degree of versatility equal to that of the space probe missiles themselves.

Research and development *use* is another kind of required versatility. Lunar shots require timing to within minutes. Several weeks may pass before conditions are favorable again, so it is mandatory to launch the missile on time. A typical countdown is started hours in advance, so that lift-off can occur on schedule. This requirement means that in the countdown equipment there must be a capability to stop and go, or to wait while certain tasks are performed.

Not only must there be an automatic capability for such a program, but when necessary, a completely manual checkout and countdown capability. The operator must have visual readout at the console for all tests. He must also be provided with a recording capability so that the value of each test may be recorded and identified for later use by safety and reliability groups.

Follow-on versatility is a requirement that has recently come to light. Fifteen years ago it was hardly considered necessary, or even possible, that the support equipment be useful for later programs, possibly because the equipment was not so costly at that time. Today's space programs, however, are vast and complicated. The necessary checkout and countdown equipment is so expensive that it must be reasonably useful for succeeding programs.

Digital Checkout and Countdown Set

Fig. 2 shows the block diagram of a centrally-programmed digital checkout and countdown set as a solution to the requirements for performance reliability and for the various kinds of versatility. It is presented in system fashion only, since this level of presentation is best suited to show just how such a set meets the requirements. The system details are too many to permit a lucid explanation of the operation by referring to it

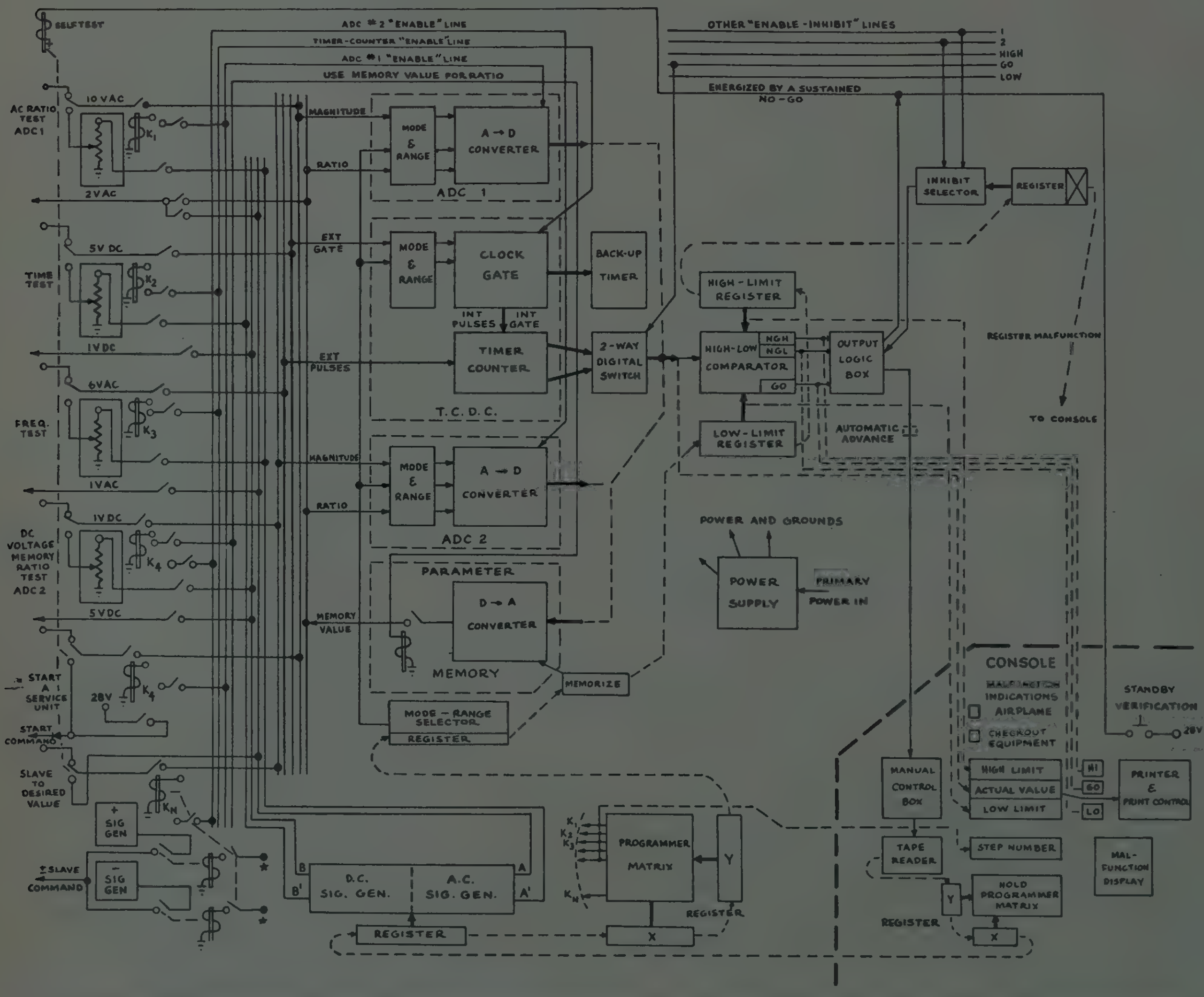
SELFTEST
ADC #2 "ENABLE" LINE
TIMER-COUNTER "ENABLE" LINE
ADC #1 "ENABLE" LINE
USE MEMORY VALUE FOR RATIO
OTHER "ENABLE-INHIBIT" LINES
1
2
HIGH
GO
LOW
ENERGIZED BY A SUSTAINED NO-GO
AC RATIO TEST ADC1
10 VAC
2VAC
K_1
TIME TEST
5V DC
1VDC
K_2
FREQ. TEST
6VAC
1VAC
K_3
DC VOLTAGE MEMORY RATIO TEST ADC2
1VDC
5VDC
K_4
START A SERVICE UNIT
28V
START COMMAND
SLAVE TO DESIRED VALUE
K_N
+ SIG GEN
− SIG GEN
± SLAVE COMMAND
MAGNITUDE
RATIO
MODE & RANGE
A → D CONVERTER
ADC 1
EXT GATE
CLOCK GATE
INT PULSES
INT GATE
EXT PULSES
TIMER COUNTER
T.C.D.C.
BACK-UP TIMER
2-WAY DIGITAL SWITCH
MAGNITUDE
RATIO
MODE & RANGE
A → D CONVERTER
ADC 2
PARAMETER
MEMORY VALUE
D → A CONVERTER
MEMORY
MEMORIZE
MODE – RANGE SELECTOR
REGISTER
B
B'
D.C. SIG. GEN.
A.C. SIG. GEN.
A
A'
REGISTER
K_1
K_2
K_3
K_N
PROGRAMMER MATRIX
Y
REGISTER
X
HIGH-LIMIT REGISTER
HIGH-LOW COMPARATOR
NGH
NGL
GO
LOW-LIMIT REGISTER
OUTPUT LOGIC BOX
AUTOMATIC ADVANCE
INHIBIT SELECTOR
REGISTER
REGISTER MALFUNCTION
TO CONSOLE
POWER AND GROUNDS
POWER SUPPLY
PRIMARY POWER IN
CONSOLE
MALFUNCTION INDICATIONS
☐ AIRPLANE
☐ CHECKOUT EQUIPMENT
STANDBY VERIFICATION
28V
MANUAL CONTROL BOX
HIGH LIMIT
ACTUAL VALUE
LOW LIMIT
HI
GO
LO
PRINTER & PRINT CONTROL
TAPE READER
STEP NUMBER
MAL-FUNCTION DISPLAY
Y
HOLD PROGRAMMER MATRIX
REGISTER
X

Fig. 2—A centrally-programmed digital checkout and countdown set. Star denotes 28 volts from low and high *enable-inhibit* lines.

directly. To explain the operation simply, portions of it are broken out in succeeding diagrams.

Fig. 3 shows the heart of the checkout and countdown unit, a comparator with high and low limit registers, an output logic box, two analog-to-digital converters with proper mode and range units connected to them, a timer-counter with proper mode and range unit and a parameter memory unit.

The analog-to-digital converter transforms the various missile analog parameters to a corresponding digital number. The comparator compares this digital number with the digital numbers stored in the lower- and upper-limit registers. If the missile value lies between these limits, the comparator registers a GO condition. The digital resolution of this system is 2^{11} bits; in other words, 2048 parts for a full-scale reading, with one bit on the basic range equal to 1 mv dc. A dual (high-low) type of comparator and the upper- and lower-limit registers solve the problem of changes of tolerance and nominal level, providing part of the answer to the requirement for electronic versatility.

The mode and range boxes for both analog-to-digital converters (ADC) and the timer-counter-to-digital converter (TCDC) are remotely programmed from an easily-changed tape, giving versatility for various modes and ranges. The ADC can be used in a dc mode with ranges of 0 to 1, 5, 10, 20, 50, 100, and so on. It is also used in the ac mode with the same ranges. In the ratio mode, two voltages are programmed into the converter, and the value converted is the ratio of the one to the other.

The timer-counter utilizes a basic capability for counting pulses to measure several types of parameters. When measuring external time, the device counts pulses supplied by the internal clock during the gated time interval bounded by $T=0$ and the $T=t$ arrival of the external timing signal. An external frequency measurement is taken by counting external pulses for a precise time interval gated by the internal clock.

Period and remaining related parameters are similarly handled, with ranges sufficient to give a high counting accuracy, even for very large numbers. The output of

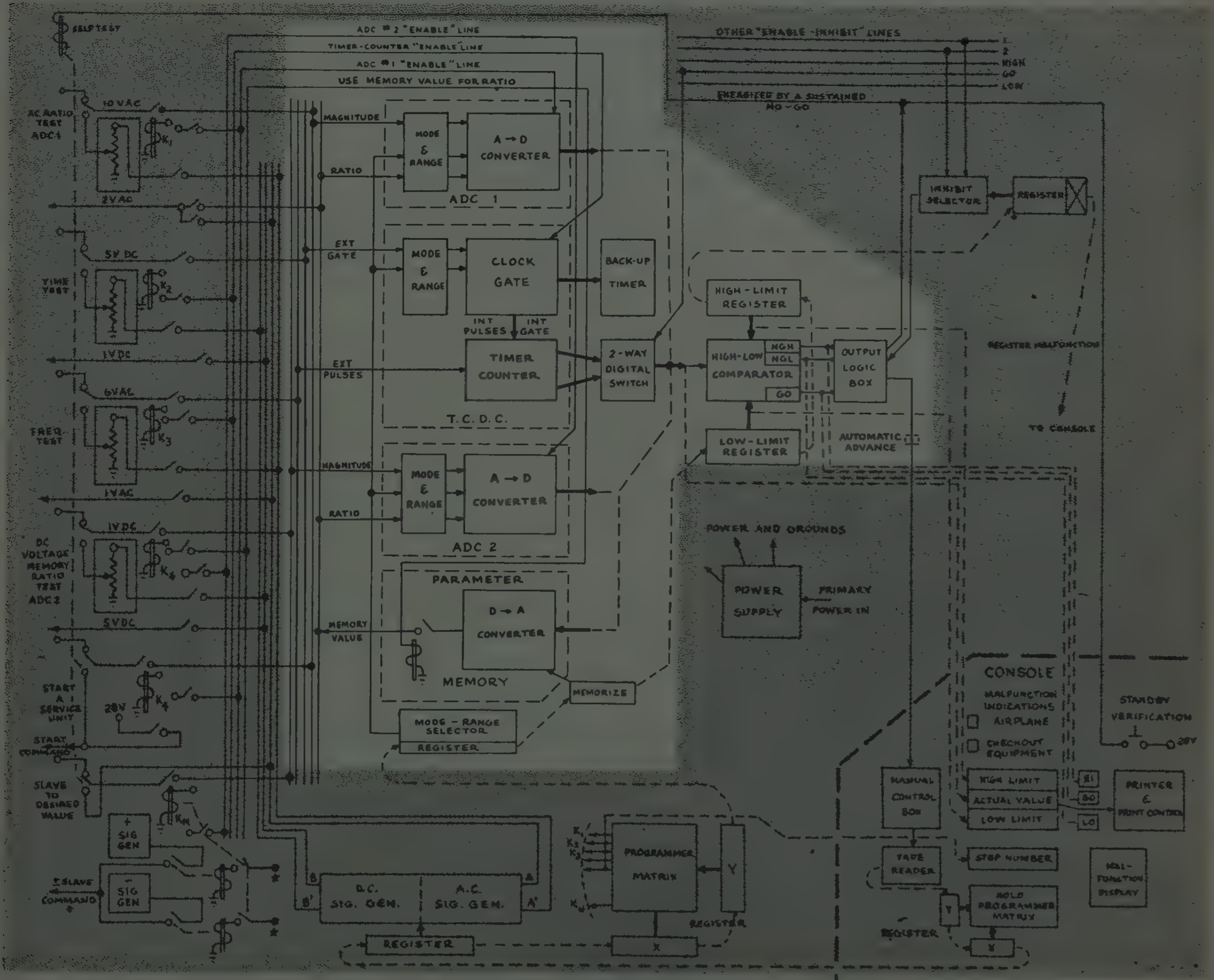
SELF TEST
ADC #2 "ENABLE" LINE
TIMER-COUNTER "ENABLE" LINE
ADC #1 "ENABLE" LINE
USE MEMORY VALUE FOR RATIO
10 VAC
AC RATIO TEST ADC 1
K_1
MAGNITUDE
RATIO
MODE & RANGE
A → D CONVERTER
ADC 1
2VAC
5V DC
TIME TEST
K_2
EXT GATE
MODE & RANGE
CLOCK GATE
BACK-UP TIMER
INT PULSES
INT GATE
1VDC
EXT PULSES
TIMER COUNTER
2-WAY DIGITAL SWITCH
T.C.D.C.
6VAC
FREQ. TEST
K_3
1VAC
MAGNITUDE
RATIO
MODE & RANGE
A → D CONVERTER
ADC 2
1VDC
DC VOLTAGE MEMORY RATIO TEST ADC 2
K_4
5VDC
PARAMETER
MEMORY VALUE
D → A CONVERTER
MEMORY
MEMORIZE
MODE – RANGE SELECTOR
REGISTER
START A SERVICE UNIT
28V
K_4
START COMMAND
SLAVE TO DESIRED VALUE
K_M
+ SIG GEN
− SIG GEN
± SLAVE COMMAND
B
B'
D.C. SIG. GEN.
A.C. SIG. GEN.
A
A'
REGISTER
K_1
K_2
K_3
K_n
PROGRAMMER MATRIX
Y
X
REGISTER
OTHER "ENABLE - INHIBIT" LINES
1
2
HIGH
GO
LOW
ENERGIZED BY A SUSTAINED NO - GO
INHIBIT SELECTOR
REGISTER
HIGH - LIMIT REGISTER
HIGH-LOW COMPARATOR
HGH
NGL
GO
OUTPUT LOGIC BOX
LOW - LIMIT REGISTER
AUTOMATIC ADVANCE
REGISTER MALFUNCTION
TO CONSOLE
POWER AND GROUNDS
POWER SUPPLY
PRIMARY POWER IN
CONSOLE
MALFUNCTION INDICATIONS
AIRPLANE
CHECKOUT EQUIPMENT
STANDBY VERIFICATION
28V
MANUAL CONTROL BOX
HIGH LIMIT
ACTUAL VALUE
LOW LIMIT
HI
GO
LO
PRINTER & PRINT CONTROL
TAPE READER
STEP NUMBER
MAL-FUNCTION DISPLAY
Y
HOLD PROGRAMMER MATRIX
REGISTER
X

Fig. 3—A centrally-programmed digital checkout and countdown set; converters, comparator, parameter memory, logic box. Star denotes 28 volts from low and high *enable-inhibit* lines.

the counter is 2^{22} digital bits. Use of the two-way digital switch allows the 11-bit comparator to evaluate a large number by evaluating first one set of lines, then the other set.

Each time the ADC-Comparator combination is used, GO and NO-GO conditions are evaluated for purposes of self-verification of the combination, for reasons identical to those noted in the explanation of Fig. 1. This is the implementation of one of the design criteria for performance reliability. A similar process of self-verification is employed when the timer-counter-comparator combination is used.

The comparator itself does not make any decisions other than LOW NO-GO, GO, or HIGH NO-GO. It supplies these outputs to the logic box, a device which decides upon the basis of predetermined logic whether the result of a test is a bona fide GO or NO-GO. If the result is GO, an AUTOMATIC ADVANCE command is given.

If the result of a particular test is a bona fide NO-GO, the logic box activates a self-test relay for the reasons explained for Fig. 1. This action implements the second criterion for performance reliability. Design-center values are automatically provided by a means to be explained in Fig. 5.

The parameter memory in Fig. 3 continually converts all digital numbers fed to the comparator into corresponding analog values. When the value of a particular parameter is to be retained for later use, a MEMORIZE command is given to the device, and the value is held indefinitely. Feeding this value into the ratio side of either ADC permits one test to be compared with another.

It is worthy of note that the parameters which can thus be compared are not limited to like dimensions, such as voltage to voltage. Just as easily, time can be compared to voltage, frequency to current, and resistance to period. Suitable ranges exist for the parameter memory, of course. They are set up by proper remote programming of the mode-range selector box, which performs a similar function for the other converters.

All the converters are permanently connected into the comparator. Each converter exists in one or the other of two states, the *inhibited* state or the *enabled* state. Only one at a time is used to feed digital information into the comparator; the remaining units are in a compatible, quiescent, inhibited state. This feature eliminates the need for a selector switch at the input to the comparator. *Enable* lines are brought out from each unit and are externally programmed to put the unit into the *enabled* state.

Various choices of ranges are to be considered. Some groups hold that decimals or powers of ten are the only ranges that should be used in a digital system. The reasoning here is that any general-purpose remote readout then can read directly in terms of the actual parameter, rather than in terms of some conditioned value of it. But the powers of ten will allow ranges of 0 to 1 volt, 0 to 10, 0 to 100, etc., only. The basic resolution is good at the maximum value of any of these ranges or just slightly below it, while at values immediately above the range it is poor.

For example, consider a missile parameter value of 9.9 volts. On the 0- to 10-volt range, the value is resolved into 999 bits—a figure adequate for most problems of tolerance. However, if the value were 11 volts, the 0- to 10-volt range would be useless and the 0- to 100-volt scale would be required. The value on this scale, instead of being divided into 1100 bits, would be divided into 110 bits out of a possible maximum of 1000 bits for 100 volts.

This situation is particularly unfortunate if a missile parameter of 11 volts must be evaluated to a tolerance of ±2 per cent, or 0.22 volt, to cite a typical problem. On the 100-volt scale, the converter has a basic resolution of 0.1 volt per bit; it would produce digital numbers corresponding to 10.7, 10.8, 10.9, 11.0, 11.1, 11.2, 11.3 volts and so on. A glance shows that the tolerance band of ±0.22 volt is resolved, or divided, into only four digital bits.

It follows that, with the best performance obtainable from a digital comparator equal to an "uncertainty" of ±1 bit, such a measurement would be unsatisfactory since the digital system is "consuming" half of the missile tolerance band. The use of powers of ten as the only choice for ranges leaves some fairly wide gaps in the resolution and evaluation capability of the countdown equipment. Fig. 4 shows how this occurs and how to alleviate the situation.

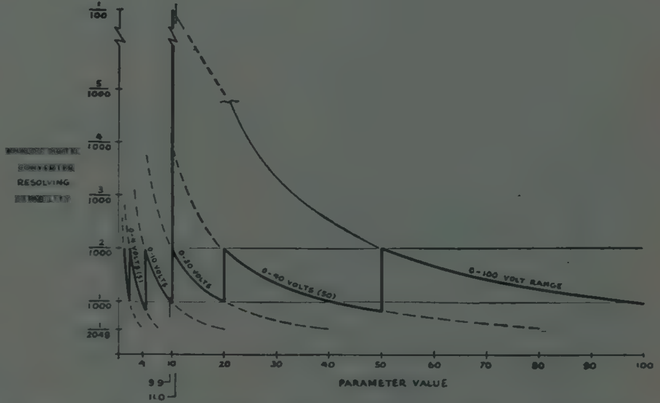

Fig. 4—Choice of multiple ranges to preserve basic evaluation accuracy. Titles for each range are *nominal* titles; for example, the 0–10-volt range may be used to 20.48 volts.

Remember that the basic accuracy and resolution of any countdown or comparison equipment should be one order of magnitude better than the closest tolerance required for any missile parameter. With the resolution equal to 1 part per 2048 for any maximum range value, it is reasonable to use as many ranges as necessary to help preserve this capability across all ranges. In this system, use is made of the basic powers of ten, plus a number of intermediate ranges. The entire set is 0 to 1, 0 to 2, 0 to 4, 0 to 10, 0 to 20, 0 to 40, 0 to 100, etc., with the 4- and 40-volt ranges used to 5 and 50 volts.

Fig. 4 shows that the worst resolution with this set of ranges is only 2 parts per thousand. To be sure, the remote readout on ranges other than the powers of ten is a readout in terms of a conditioned value rather than of the actual value. However, since the upper and lower limits are also read out in terms of the conditioned value, it is the opinion of some that visual readout of conditioned values does not present any particularly great psychological problem.

Fig. 5 shows a number of relays along one side of the diagram. One is required for each test. It connects the proper signal to the missile, connects the response from the missile to the proper signal conditioner, activates the design-center value, and supplies an *enable* command to the proper evaluator. Several of the more commonly required types of tests are shown, the most interesting one of which is the slaving relay. In this instance, the checkout and countdown unit is used as a servomechanism to accomplish such tasks as torquing a gyroscope to a desired value and rotating a tracking antenna to a position according to the numbers set into the upper- and lower-limit registers.

On the lower side is shown a random sequence programmer matrix. It accepts a digital number $K_1 \cdots K_n$ in any sequence and energizes the proper relay $K_1 \cdots K_n$. Tests or groups of tests which are not required for a particular countdown are not programmed. Additional flexibility is inherent in the fact that the racks and chassis containing relays $K_1 \cdots K_n$ are wired in a standardized fashion. The use of a new subsystem in a missile calls for a new countdown chassis with relays properly connected.

The standard signal generators are digitally programmable voltage units. According to the number in the register, they supply both an ac signal and a dc signal to line A-B. They supply ac and dc signals, which are sev-

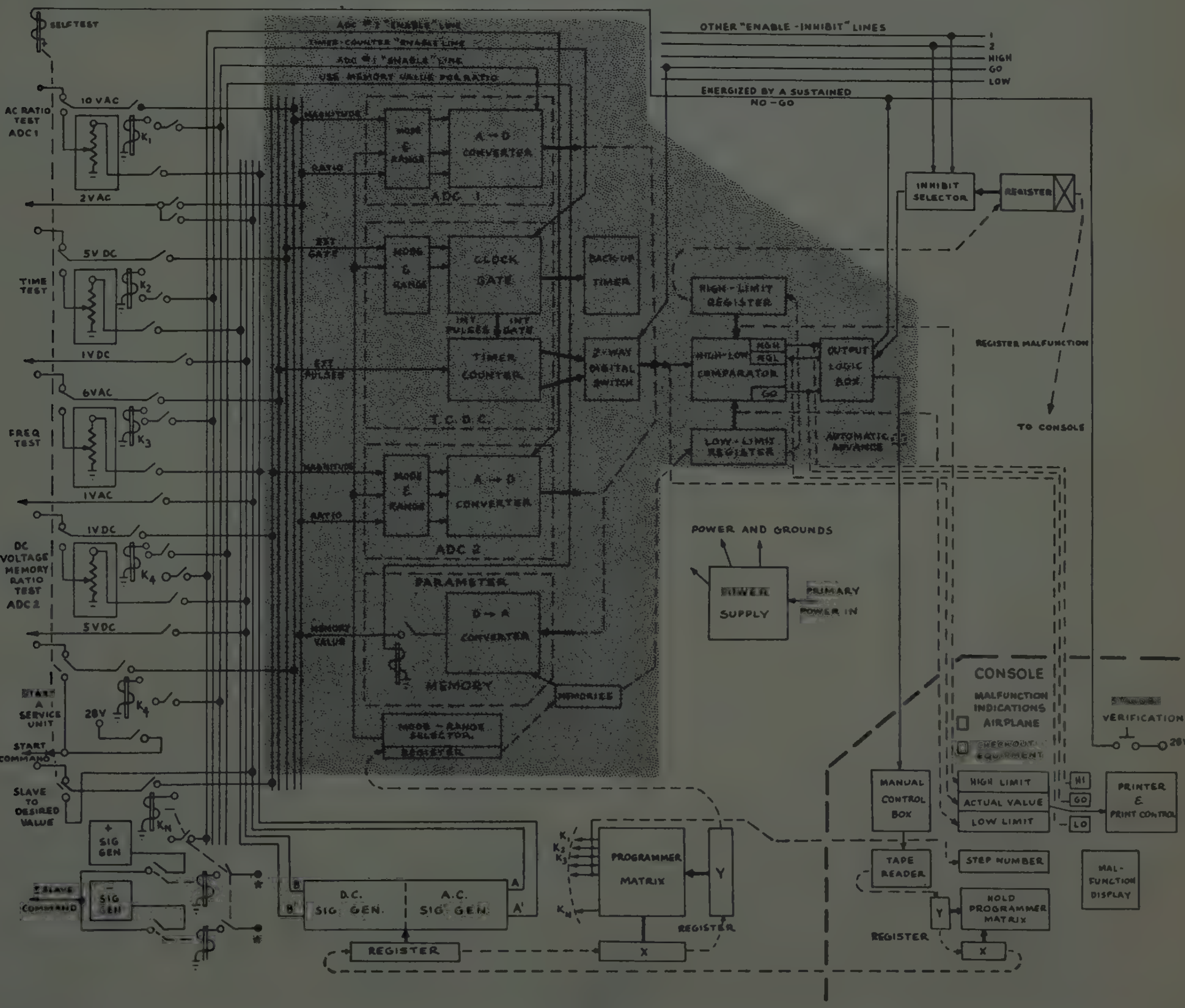
SELFTEST
ADC #2 "ENABLE" LINE
TIMER-COUNTER "ENABLE" LINE
ADC #1 "ENABLE" LINE
USE MEMORY VALUE FOR RATIO
OTHER "ENABLE-INHIBIT" LINES
1
2
HIGH
GO
LOW
ENERGIZED BY A SUSTAINED NO-GO
10 VAC
AC RATIO TEST ADC 1
K1
2 VAC
MAGNITUDE
RATIO
MODE & RANGE
A → D CONVERTER
ADC 1
5V DC
TIME TEST
K2
1 V DC
EXT GATE
CLOCK GATE
BACK-UP TIMER
INT PULSES
INT GATE
EXT PULSES
TIMER COUNTER
2-WAY DIGITAL SWITCH
T. C. D. C.
6 VAC
FREQ TEST
K3
1 VAC
INHIBIT SELECTOR
REGISTER
HIGH-LIMIT REGISTER
HIGH-LOW COMPARATOR
HIGH
REL
GO
OUTPUT LOGIC BOX
LOW-LIMIT REGISTER
AUTOMATIC ADVANCE
REGISTER MALFUNCTION
TO CONSOLE
1 V DC
DC VOLTAGE MEMORY RATIO TEST ADC 2
K4
5 VDC
ADC 2
PARAMETER
MEMORY VALUE
D → A CONVERTER
MEMORY
MEMORIES
MODE-RANGE SELECTOR
REGISTER
POWER AND GROUNDS
SUPPLY
PRIMARY POWER IN
A SERVICE UNIT
28V
START COMMAND
SLAVE TO DESIRED VALUE
KN
+ SIG GEN
− SIG GEN
COMMAND
B
B'
D.C. SIG. GEN.
A.C. SIG. GEN.
A
A'
REGISTER
K1
K2
K3
KN
PROGRAMMER MATRIX
Y
REGISTER
X
CONSOLE
MALFUNCTION INDICATIONS
AIRPLANE
VERIFICATION
28V
MANUAL CONTROL BOX
HIGH LIMIT
ACTUAL VALUE
LOW LIMIT
HI
GO
LO
PRINTER & PRINT CONTROL
TAPE READER
STEP NUMBER
MAL-FUNCTION DISPLAY
Y
HOLD PROGRAMMER MATRIX
REGISTER
X

Fig. 5—A centrally-programmed digital checkout and countdown set; test relays, programmer, signal generators, console. Star denotes 28 volts from low and high *enable-inhibit* lines.

eral times greater, to lines A'-B'. The signals on lines A and B are picked up by the various relay contacts and routed to the proper parts of the missile system. The signals on lines A'-B' provide design-center values for malfunction localization.

Should a malfunction occur, the checkout equipment will positively isolate the malfunction to the missile or to the checkout and countdown equipment. Careful control of the sequence in which tests are performed permits location of the trouble within the missile to a particular malfunctioning unit without additional tests or subroutines.

The console houses the remaining portions of the set needed for performance reliability and flexibility. The operator uses the manual control panel to disable the AUTOMATIC ADVANCE command whenever it is necessary to proceed on a manual basis. The four arabic readouts of step number, low limit, actual value, and high limit afford him adequate information to evaluate visually the test results. A MANUAL ADVANCE push button is provided on the manual control panel.

Monitoring panels are not shown. Each parameter is displayed on a relay meter with adjustable high- and low-limit switches. These switches can be connected in any combination of AND and OR logic necessary, with the total monitoring function serving as an INHIBIT function to the AUTOMATIC ADVANCE signal and energizing a suitable alarm.

The hold programmer serves two purposes. One is that of scheduling a HOLD in the countdown while a planned manual routine is being carried out. Each HOLD of this type illuminates proper instructions on the console, such as PERFORM MANUAL ROUTINE A. The operator does not proceed until he receives verification over the communication system that the routine has been properly carried out.

The other purpose served by the hold programmer is that of coordinating the primary checkout and countdown set with auxiliary equipment. When sequencing between two equipments is required, a coordination hold is programmed. The set waits until a signal arrives from the other equipment, signifying that the desired test has been performed; then it proceeds. The programmer is of the random sequence type; it can be scheduled during any portion of the countdown.

The tape reader provides all the information needed in the various registers throughout the system. The tape itself may be of any type; aluminum-backed mylar is often used. A readout speed of 100 to 200 frames per second will permit an over-all rate, for the set, of three to four tests per second.

The malfunction display is a card prepared for each flight, listing by step numbers the particular portion of the missile being tested. The printer records the pertinent data of each test with the step number. A print control permits the printer operation to be inhibited.

As in Fig. 1, standby verification is provided by a manual switch to energize the self-test relays. The set may be installed in several configurations. In a typical launch pad, the bulk of the equipment can be installed in a nearby sheltered area, with the console in a safe area several hundred feet away. Lead-length problems inevitably appear on some of the missile high-impedance test points. The two converters are provided so that, if necessary, one can be installed immediately adjacent to the missile.

Conclusion

The set is versatile. Technical changes can be accommodated by preparing a new tape. The capabilities for manual operation, for programming holds as desired, and for making visual evaluations on all tests are adequate for research and development purposes. The set can be installed in several configurations. The basic programming, evaluation, readout, and recording portions are readily usable for follow-on programs.

Performance reliability is inherently high in this centrally-programmed digital checkout and countdown set. Test results are verified immediately. Malfunction localization is positive. Standby verification is simple. All three criteria are fulfilled in a simple manner. Military equipment embodying these criteria has already been designed; use of it instills a high level of confidence in the operator.

Radiation Instrumentation Electronics for the Pioneers III and IV Space Probes*

CONRAD S. JOSIAS†, MEMBER, IRE

***Summary*—One of the devices carried aboard lunar payloads, Pioneer III and Pioneer IV, was a radiation measurement instrument. Its purpose was to detect and process information on particle flux rates encountered along the nominal cis-lunar trajectory to be followed by those payloads.**

Radiation data were collected by two different Geiger-Mueller tubes. One tube registered particle counts and the resultant accumulation was stored in a 17-stage scaler. The other tube provided average pulse current that was a compressed function of the counting rate. This current was fed to a resistor and the voltage thus formed was transferred to a telemetering subcarrier oscillator by way of a stable dc amplifier.

The electronics discussed in this paper fall into three groups: 1) dc amplifier for the analog experiment, 3) high-voltage circuitry for the detectors, 3) digital circuitry for the counting experiment.

Details of the encoding system, which permits a large dynamic range of counting rates to be clearly presented on a single subcarrier, are also discussed.

INTRODUCTION

WHEN Pioneer IV successfully escaped the earth's gravitational pull to pass in the vicinity of the moon, it carried an instrument intended to measure the radiation intensity along a nominal trajectory between the earth and the moon and possibly beyond. Its Geiger-Mueller detectors registered charged particles having energies above certain thresholds determined by detector wall thicknesses. The detectors were also capable of low-efficiency gamma radiation measurements. Cognizance for this radiation experiment was held by Van Allen of the State University of Iowa (SUI), whose responsibility entailed definition of the experiment, calibration of the flight packages and interpretation of flight results.[1,2] In addition, he specified the radiation detectors for the experiment and outlined, in general terms, the type of system to be used with the detectors. The design and construction of the electronic "radiation" package containing counters, pulse circuitry, a high-voltage supply, dc amplifiers, and data coding and compression devices were done at the Jet Propulsion Laboratory (JPL), California Institute of Technology, Pasadena, California.[3]

An earlier attempt was made by the Army Ballistic Missile Agency (ABMA) and JPL to inject a payload into a lunar trajectory. This first cone-shaped assembly, Pioneer III, did not achieve escape velocity and as a result soared to an altitude of 63,500 miles before returning to the earth. Ironically, measurement of the earth-bound radiation fields was more complete than if the probe had achieved its lunar objective. Extensive data were compiled not only on the outward leg of the payload trajectory but also on the return phase. The results of this experiment indicated for the first time the depth and contour of the partially explored radiation belt. Two Geiger-Mueller tubes were used aboard Pioneer III as radiation detectors. One of the tubes was used as a standard particle counter whereas the other tube was used as an indicator of average Geiger-pulse current. A package of nearly identical design was flown aboard Pioneer IV. The only modification consisted of 4 grams/cm^2 of lead shielding around the current averaging Geiger tube for the purpose of providing additional radiation energy data.

OUTLINE OF CIRCUIT OPERATION

As indicated in the block diagram of Fig. 1, the radia-

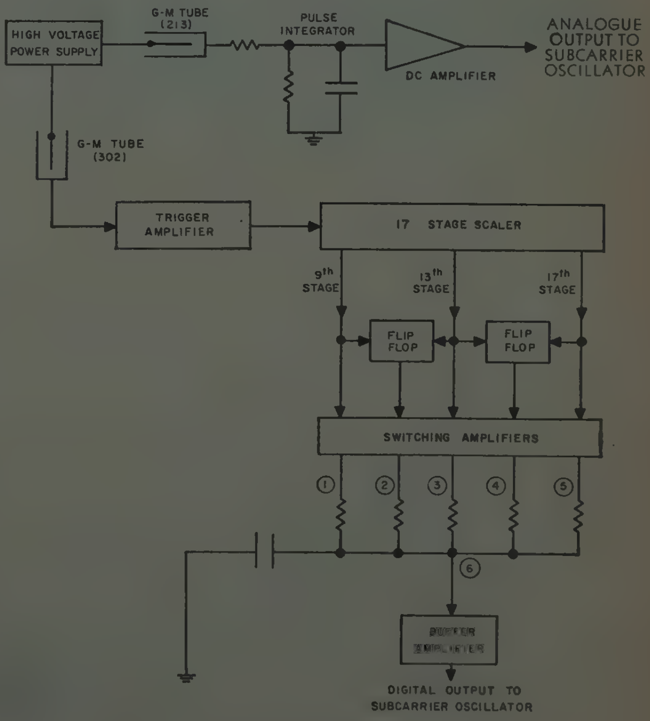

Fig. 1—Block diagram of radiation experiment.

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* Original manuscript received by the IRE, December 2, 1959. This paper presents the results of one phase of research carried out at the Jet Propulsion Lab., Calif. Inst. Tech., under Contract No. NASw-6, sponsored by the Natl. Aeronautics and Space Admin.

† Jet Propulsion Lab., California Institute of Technology, Pasadena, Calif.

[1] J. A. Van Allen and L. A. Frank, "Radiation around the earth to a radial distance of 107,400 km," *Nature*, vol. 183, pp. 430–434; February 14, 1959.

[2] J. A. Van Allen and L. A. Frank, "Radiation measurements to 658,300 km, with Pioneer IV," *Nature*, vol. 184, pp. 219–224; July 25, 1959.

[3] C. S. Josias, "Pioneer's radiation detection instrument," *Astronautics*, vol. 4, pp. 32–33, 114–115; July, 1959.

tion experiment was divided into two measurements. In one measurement, pulses obtained from an Anton Type 302 Halogen-quenched tube are processed through a trigger amplifier and then counted in a scaler from which multiple digital outputs are mixed and fed to a subcarrier oscillator. Current pulses obtained from an Anton Type 213 Geiger tube are integrated in an RC filter and the average voltage thus formed is fed to a second subcarrier oscillator through a high-input-impedance unity-voltage-gain power amplifier. The resultant voltage change at the output of the amplifier or frequency shift of the subcarrier oscillator is a compressed analog representation of the radiation intensity. The circuitry just described consumed 250 milliwatts of power at nominal battery voltage (7.2 volts) and is contained in the assembly shown in Fig. 2.

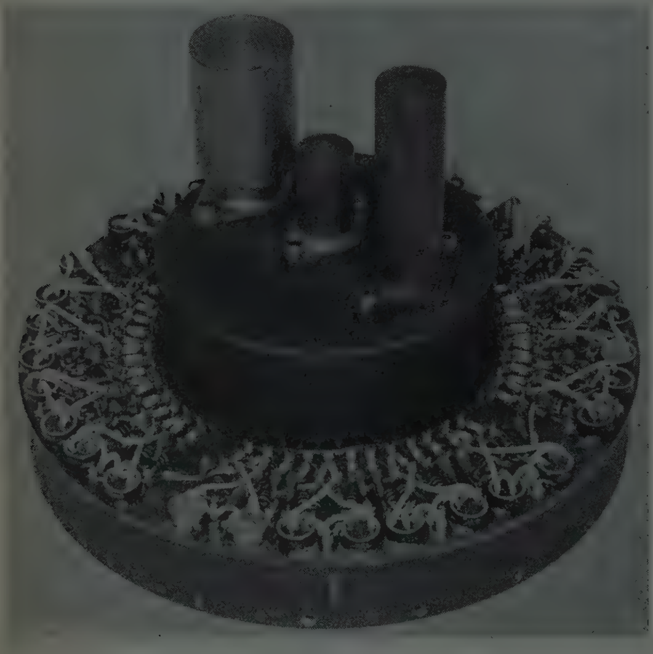

Fig. 2—Pioneer IV subassembly, top view.

DC Amplifier

One of the occupational hazards confronting the modern scientific experimenter is the continual materialization of information which often makes even new systems obsolescent. Information of this sort did indeed materialize, literally out of thin air (or space), late in the instrument development phase. Results of radiation experiments conducted by Van Allen in the Explorer IV earth satellite were observed a few weeks before scheduled delivery of the prototype radiation subassembly to the JPL Payload Group. The radiation intensity at apogee was still increasing with altitude and the peak was nowhere in sight. Another Explorer IV measurement, similar to that being prepared in the lunar package, contained a lead shielded Anton Type 302 tube. This tube registered only a modest reduction of particle counts. A new detector, the Type 213 tube, capable of indicating substantially higher radiation intensities, was substituted for the lead shielded 302 tube. A low drift electrometer-type circuit was designed to convert maximum average currents from the 213 tube into output currents having magnitudes large enough to supply signal current requirements for the associated subcarrier oscillator. Because of weight, schedule, and communications considerations, it seemed most prudent then to use an unstabilized, yet low-drift, circuit such as the one described below.

Total drift allocated to the measurement was 4 per cent of the maximum expected output excursion. Of this figure, 3 per cent was awarded to the subcarrier and 1 per cent to the amplifier. Since the subcarrier oscillator was subject to supply-voltage induced frequency shifts its supply was regulated. The remaining portion of drift could be attributed chiefly to temperature effects, and, as a result, painstaking payload temperature calibrations made it possible to reclaim much of the error that might be assigned to temperature changes.

An important aspect of the amplifier design is the dynamic range problem. Large input signals generally minimize the voltage drifts of the input differential comparator and the subcarrier oscillator. This requires that a large voltage-forming resistor be connected at the input. Current drift effects in the input stage can be reduced by inserting a similar resistor in the feedback loop. The maximum amplifier swing is limited, however, by the usable range of voltages obtainable from the payload batteries, namely 8 volts at the outset of the experiment to 6.6 volts after about 90 hours. Referring to the diagram of Fig. 3, it can be seen that if the emitters of Q1 and Q2 were returned to ground through a resistor and the emitters were expected to operate anywhere between the supply voltage and ground, the amplifier drifts due to common-mode effects would be fearsome to contemplate. A semistable constant-current source in the form of Q3 is provided for this reason. A minimum collector voltage of one volt for Q3 establishes the rest potential of the amplifier output at about 1.6 volts. The maximum output swing obtainable with a 6.6 volt supply is about 6.6 volts limited by saturation of Q1 and Q6. Of the available 5-volt dynamic range, a

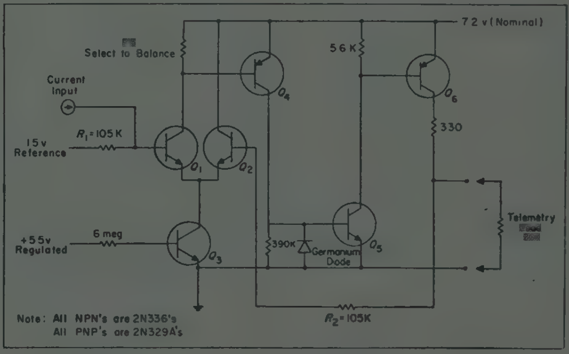

Fig. 3—DC amplifier.

maximum swing of 4.2 volts was selected for measuring currents up to 40 μa. The voltage-forming resistance for these values of current and voltage is 105K.

Three factors which might contribute to amplifier voltage drift that can be assigned to the input differential comparator are

1) imperfect tracking of base-to-emitter voltages in Q1 and Q2;
2) changes in common-mode or average input currents to Q1 and Q2; and
3) changes in differential input currents to Q1 and Q2.

Under normal operating conditions ($V_{be}=0.5$ volt), the temperature coefficient of base-to-emitter voltage for silicon transistors is about −2 mv/°C at room temperature.

To insure tracking of these characteristics in Q1 and Q2, both units were selected with matching transfer characteristics (*i.e.*, V_{be} vs I_c) and, as an added precaution, similar current gains at collector currents of about 10 μa. The final amplifier drift adjustment consisted of balancing the operating points by proper selection of R3. Surprisingly, the β of the 2N336, when operated at starved collector currents of about 10 μa, was still a respectable 25–30 at room temperature.

To study the effects of current drifts in the differential comparator, consider the equivalent circuit in Fig. 4. In this diagram, the voltage-forming resistors connected to the base of Q1 have been replaced by their equivalent source resistance and voltage as observed from the base of Q1. If, for some reason, changes occur in the dc bias currents I_1 and I_2, the circuit voltage changes may be shown as

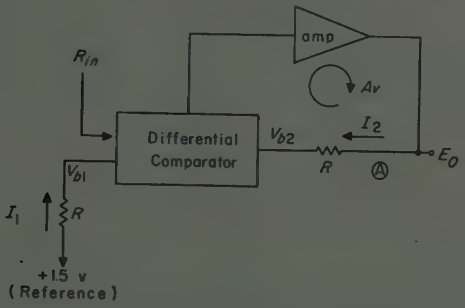

Fig. 4—DC amplifier—equivalent circuit.

$$\Delta V_{b1} = -R\Delta I_1$$

and

$$\Delta V_{b2} = \Delta V_{b1} - (\Delta I_1 - \Delta I_2)R_{bb}$$

where R_{bb} is the base-to-base resistance. As a result, the change in output voltage may be written

$$\Delta E_0 = -\Delta I_1 R - (\Delta I_1 - \Delta I_2)R_{bb} + \Delta I_2 R,$$

$$\Delta E_0 = (\Delta I_2 - \Delta I_1)(R + R_{bb}).$$

The drift effect noted here is due to changes in differential base current. The value of R was 105K and a representative value of R_{bb} at the appropriate starved operating point is 150K. An output drift of 10 mv would therefore be produced by a differential current change as small as 0.04 μa. One of the mechanisms causing temperature-induced drift is change of gain in Q3. As temperature increases, the gain of Q3 increases and its collector current is correspondingly higher. Virtually all of the increase in current is absorbed by Q2. Here is a case in which differential base currents to Q1 and Q2 change and base-to-emitter voltages also become unequal. A subsidiary but nonetheless interesting corollary of unbalanced base-to-emitter voltages is that the respective emitter diodes have unequal temperature coefficients of voltage. The tracking error between two transistors having voltages of 0.4 and 0.5 volt, respectively, is about 0.33 mv/°C. To ease the multiple maladies that occurred with changing temperature, a germanium diode was inserted in the base of Q4. As temperature rose the increased I_{co} drawn by the diode was reflected back to Q1, which was now better able to absorb its share of the increased current generated by Q3. Temperature performance of a representative amplifier is shown in Table I.

TABLE I

TEMPERATURE PERFORMANCE OF DC AMPLIFIER

Temperature	Input Current (μa)	V supply =6.6 volts	V supply =7.2 volts	V supply =8.5 volts
+75°C	0	1.637	1.637	1.635
	20	3.768	3.767	3.766
↓	40	5.885	5.884	5.883
+50°C	0	1.624	1.623	1.622
	20	3.758	3.757	3.755
↓	40	5.867	5.865	5.862
+25°C	0	1.605	1.605	1.605
	20	3.731	3.730	3.730
↓	40	5.843	5.842	5.840
0°C	0	1.579	1.579	1.577
	20	3.713	3.713	3.712
↓	40	5.815	5.814	5.812
−10°C	0	1.570	1.569	1.568
	20	3.696	3.696	3.695
↓	40	5.805	5.804	5.802

To insure circuit protection during early phases of payload system integration, a protective resistor was inserted in the collector of Q6. If the output is inadvertently shorted to ground, the 1-μa bias current in Q3 is amplified with full open-loop current gain thereby forcing huge currents through Q6 and the short. The 330-ohm resistor in the collector of Q6 limits current to a safe maximum and provides simple short-circuit protection.

The resistance seen looking into the base of Q1 is extremely high and can be approximated by the expression

$$R_{\text{in}} = (R + R_{bb})(1 + Av).$$

The open loop gain, Av, may be measured by breaking the loop at point A as indicated in Fig. 4 (with the input base at ac ground). The high input resistance is produced by the same bootstrap effect that reduces input capacitance to a cathode follower. This resistance is high enough so that one can no longer neglect the base-to-collector resistance of Q1. In any event, the loading on the voltage forming resistor is negligible.

The thermal operating point of Pioneers III and IV was +40°C. The drift coefficient of the amplifier described in Table I in the 40°C region was about 0.75 mv/°C.

HIGH-VOLTAGE POWER SUPPLY

The high-voltage power supply for the Geiger tubes contains a transistor-saturable-core oscillator which runs directly off the payload batteries and supplies square-wave power to a rectifier and voltage-doubler circuit. DC regulation is accomplished with a 700-volt Corona regulator. The regulated outputs of the supply are 730 volts and 650 volts for the Types 302 and 213 tubes, respectively. The high-voltage circuit, which is located inside the "crown" of the subassembly directly beneath the Geiger tubes (see Fig. 5), was foam-potted to protect against high-altitude corona discharge.

Corona-tube idling current was selected to be a value larger than the largest expected combined Geiger-tube currents plus the minimum Corona-tube operating current. An initial idling current appreciably larger than the maximum expected load permits the regulator to handle these demands at the end of the experiment when battery voltage is low and idling current has been significantly reduced. If, for instance, maximum output were expected of both Geiger tubes near the end of the experiment (an extremely remote possibility in the vicinity of the moon), currents of about 50 μa would need to be handled by the regulator. Since the high-voltage power supply was to dissipate the major portion of power supplied to the radiation subassembly, care was required in selecting the lowest power mode of operation consistent with performance requirements. This meant that the transformer turns ratio had to be specified to provide unregulated high-voltage at minimum average power levels over the duration of the experiment. Consider the power-handling problem presented by the regulator shown in Fig. 6, which is similar to the actual regulator in the system schematic shown in Fig. 7. The power delivered by the inverter at the beginning of the experiment is $P_1 = E_1 I_1$, where E_1 is the unregulated voltage at the outset and I_1 is the regulator current expressed as $(E_1 - 700)/(R_1)$. At the end of the experiment, when the battery voltage has dropped to 6.6 volts, indicating that it has reached the limit of its useful life, the unregulated power delivered by the inverter has dropped to $P_2 = E_2 I_2$. The voltage E_1 may be written

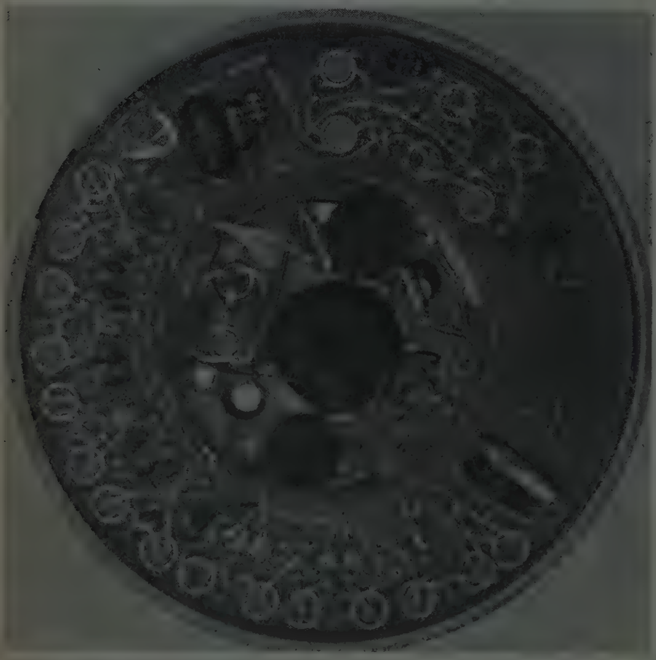

Fig. 5—Pioneer III subassembly, bottom view (power supply unfoamed).

$$E_1 = \frac{8.0}{6.6} E_2 = 1.2E_2.$$

The power delivered at the beginning of the experiment in terms of E_2 and I_2 is

$$P_1 = 1.2E_2 I_2 \frac{(1.2E_2 - 700)}{(E_2 - 700)} \cdot$$

Since the battery voltage decays linearly with time, the unregulated power supplied by the inverter (see Fig. 5) falls off parabolically according to the equation

$$P_{\text{inverter}} = \frac{E(E - 700)}{R_1} \cdot$$

The power points determined by E_1 and E_2 are close enough together on this parabola that the average power delivered by the inverter over the life of the experiment may be accurately expressed by the first order approximation

$$P_{\text{average}} = \frac{(P_1 + P_2)}{2} = \frac{E_2 I_2}{2}\left[1 + \frac{1.2(1.2E_2 - 700)}{(E_2 - 700)}\right].$$

The value of E_2 yielding minimum average power is determined by differentiating the expression for P_{average} with respect to E_2 and setting this derivative equal to zero. The values of E_2 and E_1 are found to be 923 and

1110 volts, respectively. The nominal turns ratio for the transformer in this example, is 70 for the case of a lossless inverter system followed by a voltage doubler. Before designing the Pioneer high-voltage transformer, the transformer used in the historic Explorer radiation experiments[4] was tested in the inverter configuration shown in Fig. 7. A new winding specification was generated on the basis of existing estimates on payload battery voltage limits, desired unregulated output voltages, and loss measurements made with the Explorer transformer. The flight-model high-voltage power supply used a transformer having a Supermalloy core and a 90:1 step-up turns ratio. The inverter oscillated at about 2.5 kc and supplied an unregulated dc output of 1000 volts at 75 μa of idling current when the payload batteries were at 6.6 volts. Under this condition, the supply efficiency was close to 50 per cent, the majority of the extra power being dissipated in core magnetization.

The Geiger tubes used in this experiment generally drew peak ionization currents well in excess of their maximum expected average current. The filter connected to the corona tube shown in Fig. 6 limits current surges through R_2 to a small fraction of the maximum average ionization current.

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[4] G. H. Ludwig, "Cosmic-ray instrumentation in the first U. S earth satellite," *Rev. Sci. Instruments*, vol. 30, pp. 223–229; April 1959.

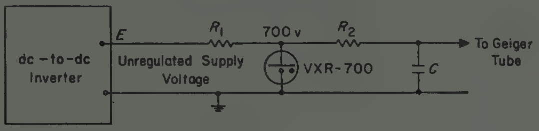
dc -to-dc
Inverter
E
R_1
700 v
R_2
Unregulated Supply
Voltage
VXR-700
C
To Geiger
Tube

Fig. 6—High-voltage regulator.

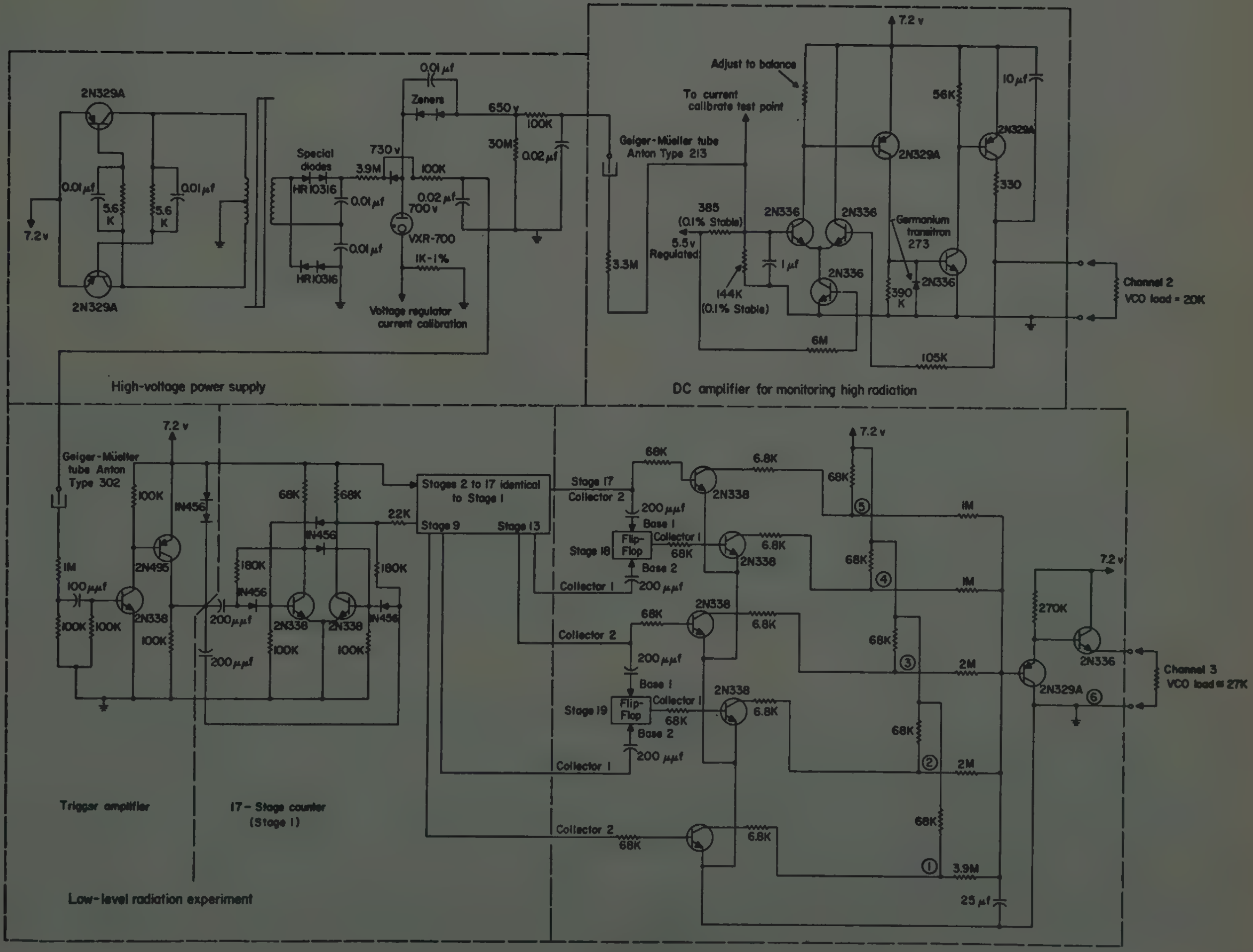
2N329A
0.01 μf
5.6 K
7.2 v
2N329A
Special diodes
HR10316
730 v
3.9M
100K
0.01 μf
0.02 μf
700 v
VXR-700
0.01 μf
HR10316
1K-1%
Voltage regulator current calibration
0.01 μf
Zeners
650 v
100K
30M
0.02 μf
High-voltage power supply
Adjust to balance
To current calibrate test point
Geiger-Müeller tube Anton Type 213
3.3M
385 (0.1% Stable)
5.5 v Regulated
144K (0.1% Stable)
1 μf
2N336
2N336
2N336
2N329A
56K
10 μf
2N329A
330
Germanium transitron 273
2N336
390 K
6M
105K
Channel 2
VCO load = 20K
DC amplifier for monitoring high radiation
7.2 v
Geiger-Müeller tube Anton Type 302
1M
100 μμf
100K
100K
2N338
100K
2N495
100K
1N456
200 μμf
200 μμf
68K
68K
1N456
180K
1N456
2N338
2N338
100K
100K
180K
1N456
22K
Stages 2 to 17 identical to Stage 1
Stage 9
Stage 13
Trigger amplifier
17- Stage counter (Stage 1)
Low-level radiation experiment
Stage 17 Collector 2
68K
2N338
6.8K
200 μμf
Base 1
Stage 18
Flip-Flop
Collector 1
68K
Base 2
Collector 1
200 μμf
6.8K
2N338
Collector 2
68K
2N338
6.8K
200 μμf
Base 1
Stage 19
Flip-Flop
Collector 1
68K
Base 2
Collector 1
200 μμf
2N338
6.8K
Collector 2
68K
6.8K
68K
⑤
1M
68K
④
1M
68K
③
2M
68K
②
2M
68K
①
3.9M
25 μf
270K
2N336
2N329A
⑥
Channel 3
VCO load = 27K

Fig. 7—Schematic: radiation experiment, Pioneers III and IV.

DIGITAL EXPERIMENT

Pulse-Forming Circuits

The circuitry connected to the shell of the Type 302 tube consists of an ionization-current-limiting resistor and an RC network coupled to the trigger amplifier (see Fig. 7). The diodes at the output of the trigger amplifier establish standard pulse amplitudes of about 1.2 volts. The input capacitor and emitter diode of the first transistor in the trigger amplifier produce a dc restoration effect similar to the one required for television video signals. The positive pulse signal has its peak restored to +0.6 volt by the transistor base-to-emitter diode. Since base current flows only during the first 10 or 15 μsec after the ionizing event and is induced only by the ionization signal peak, there is no danger of trigger amplifier saturation for the closest spaced pulses.

Counting Range

The maximum average rate at which the Type 302 tube will count, as determined by tube dead time, is about 16,000 to 18,000 counts per second. Dead time is that period of insensitivity in the tube during which no other ionizing events can be detected. At the end of this time, the field encircling the center conductor of the tube has reached its threshold value. The recovery time is the additional increment needed to restore the second pulse to its full height.[5] Both effects are illustrated in Fig. 8. When particle counting rates are far in excess of the maximum counting ability of the tube, an increasing percentage of the particles will either not be counted or will produce ionizations occurring at intervals slightly larger than the dead time. Peak-to-peak amplitude of the closely spaced pulses will be sharply attenuated resulting in a smaller percentage of the pulses being large enough to trigger the input stage of the scaler. The reduced pulse amplitude has the effect of producing a double-valued calibration curve. In the region where the apparent count rate reaches a peak and then drops off, the analog channel produces an output that still varies monotonically with radiation intensity. Therefore, the pulse integrator circuit used as a high-intensity monitor also resolves ambiguities in the digital channel caused by the double-valued nature of its response. Maximum average count rates obtainable in Pioneers III and IV were about 12,000 counts per second.

Scaler Chain

The basic flip-flop (see Fig. 9) used in the scaler chain is a saturating transistor device capable of operation up to 100,000 counts per second with fixed amplitude triggers. The flip-flop performs with "semistarved" collector currents of 100 μa per collector, dissipating 1.4 milliwatts per stage. This circuit is by no means fast, in terms of modern high-speed computer devices, but it is more than adequate for the Geiger tube counting rates encountered in this experiment. In addition, it is an unquestionably reliable tool in terms of most missile or space vehicle environments. A noteworthy aspect of the flip-flop performance is its ability to operate on, and also deliver, trigger pulses of relatively constant amplitude irrespective of supply voltage. To a certain extent this is true of most direct-coupled transistor flip-flops. The cross-coupling diodes abetted by the conventional silicon transistor base-operating potentials standardize the output swing at 1.1 to 1.2 volts. The diodes are also instrumental in promoting fast regeneration for flip-flops operating at low currents with the associated degraded transistor characteristics. "Turn-off" of a saturated device, for instance, is aided by the storage properties of the diode which replace corresponding properties of the conventional cross-coupling resistor and its charged parallel speed-up capacitor.

The steering diodes provide 1.1 volts of positive trigger rejection to the ON transistor which prevents double triggering. The operating point of the OFF transistor base is such as to provide at least 2 μa of I_{co} protection before multiplication takes place.

The entire chain of 17 flip-flops plus the two auxiliary stages were demonstrated to operate reliably between the relatively wide supply-voltage limits of 2 and 20 volts at 25°C. Similar performance was obtained between 5 and 10 volts between ambient temperature extremes of −15°C and +75°C.

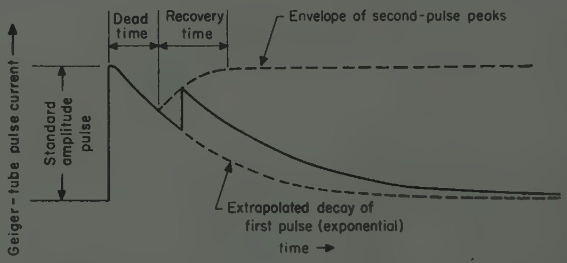

Fig. 8—Reduction in amplitude caused by recovery-time effects.

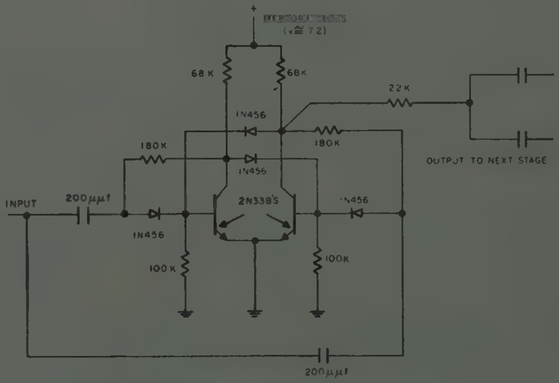

Fig. 9—Basic scaler flip-flop circuit.

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[5] S. A. Korff, "Electron and Nuclear Counters," D. Van Nostrand Company, Inc., New York, N. Y., ch. 4, pp. 130–131; 1955.

Coding

To improve the readability of the scaler output caused by a pulse input having a widely varying counting rate, three outputs were selected at four-stage intervals. These signals are mixed in the ratio of 1:2:4 for the first, second, and last tap, respectively. The result is the consolidation into one subcarrier of information that might ordinarily require three subcarriers. Reading speeds of slow counting rates are thereby improved by as much as 256:1 over that obtained from a single tap at the end of an identical number of counter stages. As a corollary, this coding process also minimizes the number of variable speed playbacks of the telemetry magnetic tapes needed for clear, readable information. This unfiltered composite waveform is shown in Fig. 10. The uncompensated mixed waveform produces a large voltage step once every 2^{17} pulses or once every major scaler cycle. To preclude the occurrence of correspondingly large step changes in frequency at the output of the subcarrier oscillator, an auxiliary flip-flop (stage 18 in Fig. 7) provides step reduction at the critical point. Similar compensation is performed on the subcycle by stage 19. The step reduction mechanism permits minimization of the output shaping-filter time constant for given values of subcarrier band swing. Actual radiation telemetering data received on the re-entry leg of the Pioneer III trajectory is illustrated in Fig. 11.

Scaler Length

The scaler chain was made long enough to prevent the fastest expected output signal appearing at the last stage from being severely attenuated by the shaping filter in the mixing circuit. By shaping the output square-waves into exponentially rising and falling edges, steady-state signal phase error occurring in the microlock receiver subcarrier phase-locked loop could be fixed at a tolerable maximum. In broad terms, the scaler length is indirectly determined by such interrelated items as sideband power, receiver loop-bandwidth, and signal-to-noise ratios at threshold distances.

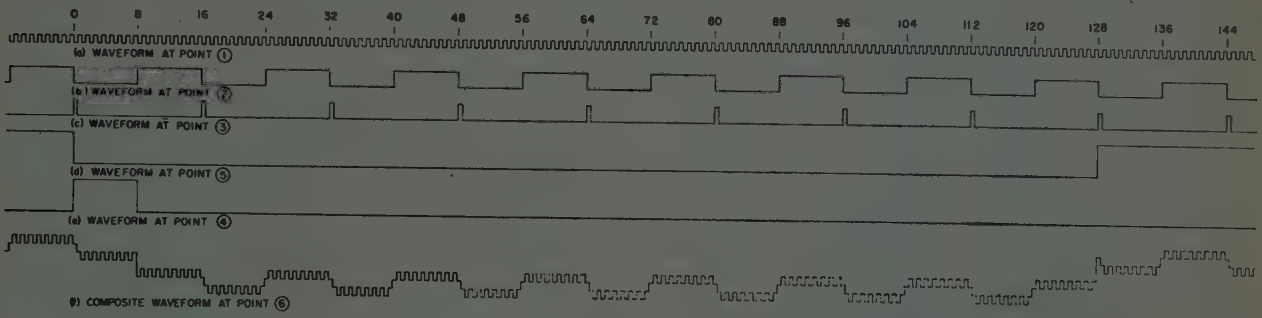

Fig. 10—Individual and composite digital waveforms. Amplitude of waveforms in (a)-(e) are illustrated according to relative weighting in the mixer circuit. Waveform locations are referenced to points in the schematic and block diagram.

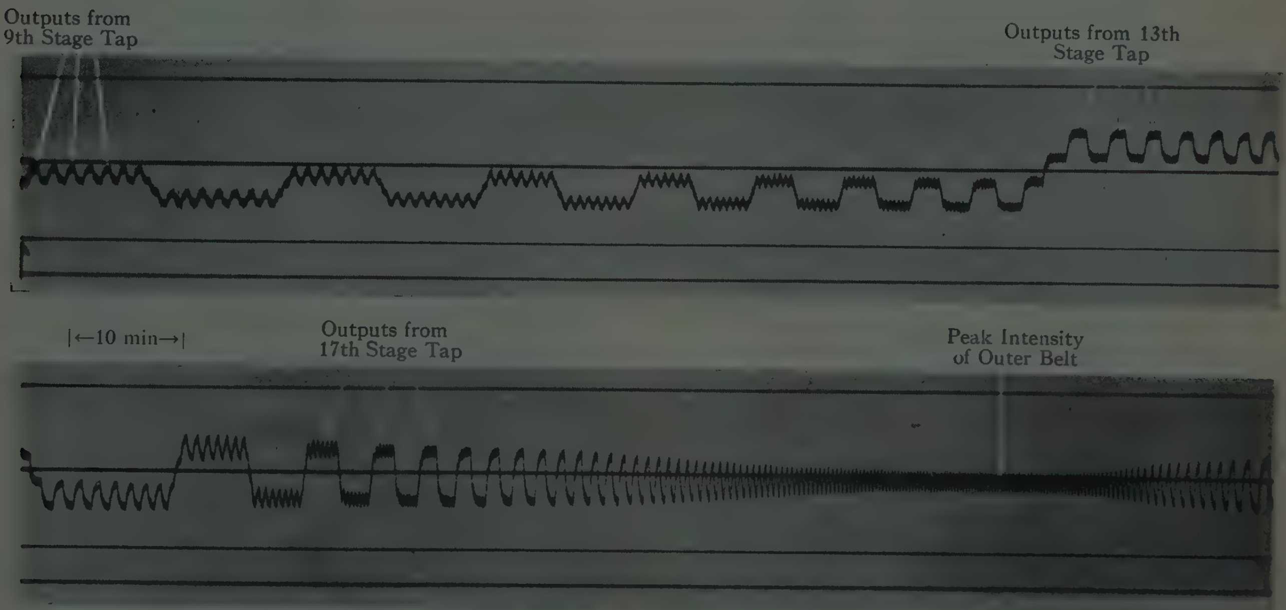

Fig. 11—Pioneer III re-entry radiation data monitored at Mayaguez, Puerto Rico (subcarrier frequency vs time).

COMPONENT CONTROL

The corona regulator used in the high-voltage power supply was the Victoreen VXR-700; the same type flown in the early Explorers. The volt-ampere characteristics of all incoming tubes were measured at the nominal operating temperature and also at the ambient extremes. All high-voltage rectifier diodes were inspected for reverse leakage currents at their rated peak inverse voltages over the same range of operating temperatures as used in testing the regulator tubes. Similarly, the low voltage Zener Diodes, used to adjust the high voltages applied to the Geiger tubes, were thoroughly tested.

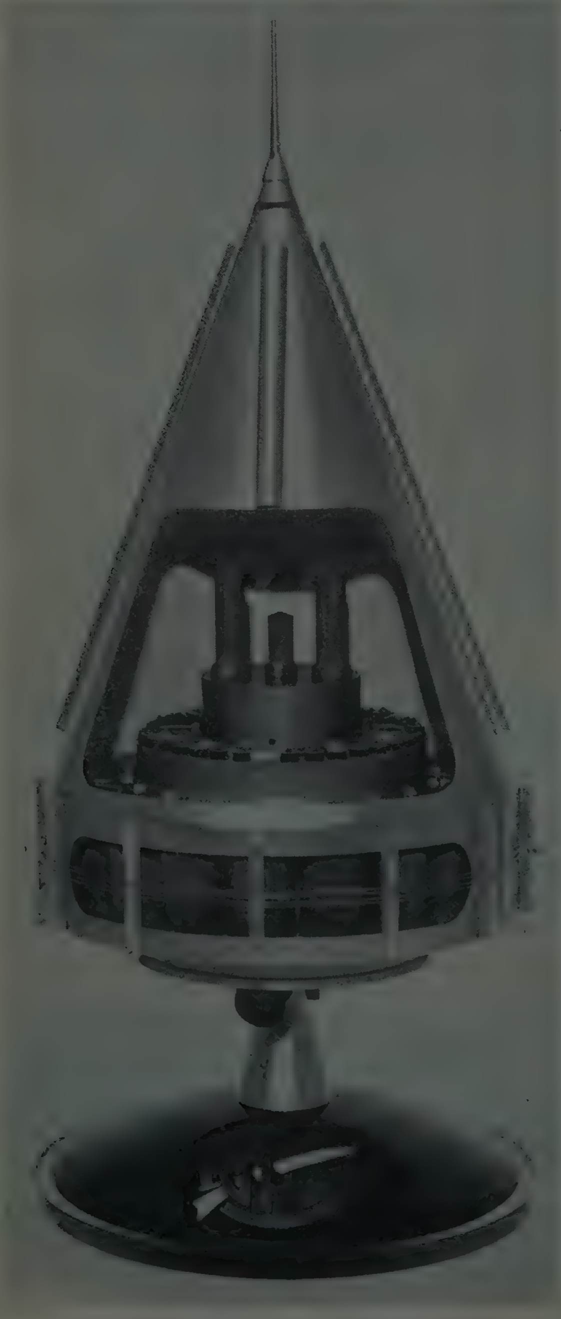

Fig. 12—Cutaway model of Pioneer III.

Part of the cooperative effort between SUI and JPL involved precalibration of all flight-approved Geiger tubes at SUI. Four final tests were performed on the Geiger tubes after they were installed in the subassembly. Before the internal high-voltage connections were made to the tubes and before the supply and Geiger tubes were foamed into place, threshold or starting voltage measurements were made on both the Type 302 and 213 tubes. A plateau measurement was then run on the 302 tube. After the high-voltage connections were completed, radiation-induced ionization pulses were inspected for height and dead time across the current limiting resistors. Typical values of dead time for the 302 tubes were 60 μsec. The maximum linear counting rate is less than 1/(dead time) because of statistical particle pile-up effects.

All transistors used in the scaler flip-flop circuits were inspected for switching current gains. The switching gain in this application was defined as the dc value of β at the 100-μa knee of the common-emitter collector characteristic. Transient switching characteristics of these transistors were also logged. All 2N336 transistors

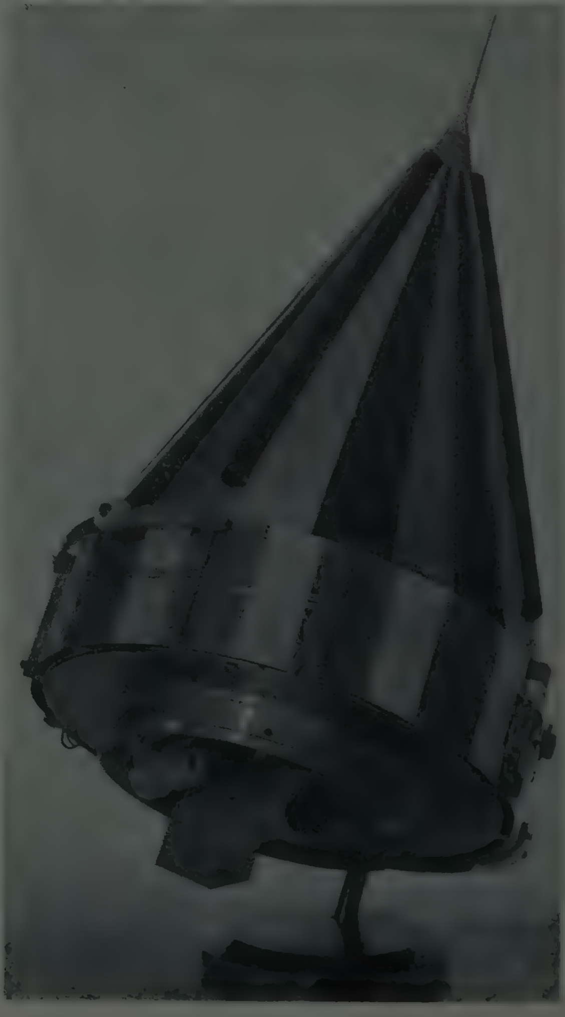

Fig. 13—Pioneer IV.

used in the dc amplifier were checked at 2 collector-current points for dc β and base-to-emitter voltage for subsequent matching.

ENVIRONMENT

A total of seven subassemblies was constructed during the course of the project. The first two were slated for type-approval tests and met the most severe set of environmental conditions that flight units might encounter. The next five were designated as flight hardware. The flight-acceptance environment was not quite so severe as the type approval in order that each flight unit might have a better flight-life expectancy.

The following flight-acceptance tests were performed on the subassembly by itself and again on the payload system, which is shown in two forms in Figs. 12 and 13.

1) Temperature: −10°C to +75°C (subassembly)
+25°C to +45°C (payload)
2) Acceleration: a) Flight Plane: 65 g for 1 minute
b) Spin: 750 rpm for 5 minutes
3) Altitude: 150 microns of Hg
4) Vibration: 10 g rms white noise between 20–1500 cps with 25 g bursts (0.2 second duration)
5) Shock: 100 g
6) Despin: 750 rpm to 6 rpm

CONCLUSION

This paper has described an electronic device characteristic of first generation scientific space instruments. Perhaps the device itself is only of academic interest in the instrument field although the scientific results, such as the radiation data transmitted to the JPL downrange station in Puerto Rico, by the instrument, are decidedly of more than just casual interest.

ACKNOWLEDGMENT

Much credit is due D. W. Slaughter for his contributions to the digital phase of this experiment. Encoding suggestions made by W. K. Victor at the outset of the project were used in the final design. The group technicians who constructed the subassemblies and earnestly performed seemingly endless testing and evaluation deserve a large measure of gratitude for their important contributions.

The author wishes to thank Dr. Van Allen and his staff at SUI for their cooperation in the preparation of the experiments. The careful integration and handling of the Pioneer payloads by the JPL Payload Systems Group, under the direction of D. Schneiderman, contributed notably to the success of the missions.

Stability of Rotating Space Vehicles*

H. L. NEWKIRK†, W. R. HASELTINE†, AND A. V. PRATT†

***Summary*—The problem of stabilizing space vehicles is considered, and it is shown that at least for some important applications the intrinsic stabilization obtained by rotation of the entire vehicle is advantageous. Investigations which have been made of natural methods for damping nutational vibrations of such rotating vehicles are discussed in the following manner.**

A laboratory simulator is described which has been found very useful in testing the various devices suggested for nutation damping. A number of references to recent analytical studies of nutation dampers are given. One approach, a rigorous attack on a simplified arrangement, is described in some detail. Also, formulas based on an intuitive approach are derived for the design of both the pendulum and the mercury dampers. Finally, an interesting and, it is believed, entirely new basic design criterion for stable operation of coupled rotating bodies, such as a space vehicle with pendulum nutation dampers, is demonstrated.

INTRODUCTION

A NONROTATING rigid body in space possesses little or no inherent stability, since even a small torque exerted for a short time will result in some angular velocity, or tumbling, which continues until countered by an equal and opposite torque. However, a similar rigid body which is rotating has an angular momentum vector with a direction fixed in space. This direction will change as the result of a torque on the body normal to the momentum axis, but it continues to change only as long as the torque is applied and then it will stop. Spinning a space vehicle, therefore, provides a kind of intrinsic stability or resistance to disturbing torques and, although it may be quite possible to stabilize a nonrotat-vehicle by means of an independent inertial reference controlling a system of jets, stabilization based on a natural property, such as spin of the entire vehicle, should be more economical as well as more reliable.

For some space vehicles there may be functional objections to spin but for many applications spin will actually fulfill specific requirements other than stability. As an example: the large space stations are invariably envisioned as spinning slowly to provide a force field or artificial gravity in the living quarters which are located

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at a distance from the center of rotation. On a much smaller scale, present-day vehicles intended to view the moon surface or the earth's cloud cover are designed to rotate about an axis normal to that of the telescope.[1] This provides automatic scanning of the area to be viewed. There are, undoubtedly, many other applications for rotation in space vehicles and, because of the intrinsic nature of spin stabilization, thorough study of the dynamics of the spinning space vehicle is well warranted.

It was noted that the momentum axis of a spinning body would change direction under an applied torque normal to this axis, and when the torque is released the motion would stop. However, the spin axis may continue to move in a nutational vibration. Here, as in mechanics texts, the motion of the axis of rotation under such an external torque is termed "precession." During precession the total angular momentum vector of the body actually changes in direction. By contrast "nutation" describes the rotary oscillations of the spin axis of the body about the total angular momentum vector. Nutation is a vibratory type of motion like that of a mass attached to the end of a thin, spring-steel member and taking a rotary motion about the equilibrium position; similarly, nutation occurs as the result of the initial conditions being different from a stable state. Nutation can occur simultaneously with, and superimposed upon, precession, which can give rise to the familiar epicyclic or bobbing motion of a common top.

In order to change the attitude of a rotating space vehicle, probably the simplest method is to precess it by means of a pair of jets fixed in the vehicle. The jet torque must therefore operate during a portion of each revolution (for example 90°) for successive revolutions of the vehicle until the desired precessional angle has been attained. Note that the jets can be made to operate in any sector during a revolution of the vehicle so that the axis can be precessed toward any desired direction by a single jet fixed system. However, the torque causing the desired precession is not constant or even slowly varying but occurs in a series of discontinuous pulses in synchronism with the spin, and nutational vibrations can be expected from this as well as from other disturbances. For most applications the nutation is objectionable if not intolerable, and should be damped as quickly as possible to a negligible value.

As with the basic attitude control, damping could probably be accomplished with an independent inertial reference and jets. However, again there appear to be natural methods to obtain nutation damping, using passive instead of active devices.

A similar need for damping nutational vibrations has arisen in the design of various guidance components which depend on gyroscopic action. A very effective solution has been the mercury damper. In one design this consists of two circular channels in a disk mounted concentric with the axis of the gyroscope wheel. The channels are only partially filled with mercury so that it can move freely within them. As discussed in a later section, apparently the diameter of the channels in relation to the other system parameters influences the effectiveness of the damping, and the design has provided excellent damping characteristics.

However, when the mercury damper was proposed for space vehicles an objection was raised. It appears that the same (passive) device which produces damping in a disk-shaped object rotating about the axis of greatest moment of inertia will cause an undamping of nutations in a cigar-shaped vehicle rotating about its longitudinal axis. It is usually quite feasible to construct a space vehicle *payload* shaped basically like a short, flat cylinder. However, rocket boosters are conventionally long compared to their diameter and, since the firing of the last booster stage is often preceded by a long coast period after which the direction of its thrust is important, there is much concern over stability. It would be highly desirable, therefore, to have a nutation damper which could be incapacitated, or which is known to be ineffective during the boost stages, but which is extremely effective in the final space vehicle.

SPACE VEHICLE DYNAMICAL SIMULATION

Theoretical attacks on this nutation damper problem have proven very frustrating, and a simulator was designed in an attempt to investigate in the laboratory the nature of the motions with various damping devices and missile dynamical configurations. This device is shown in Fig. 1. By changing the separation of the two large rings, which are made of heavy copper, on the light aluminum center cylinder, the transverse moment of inertia, B, can be changed with respect to the polar moment of inertia, A. In this way the ratio A/B can be varied from about 0.7 to 1.55. The symmetry axis of the unit is adequately defined by the steel shaft at the top, and this also serves for attaching various experimental devices for nutation damping. The entire unit is mounted near the center of the aluminum cylinder on very low-friction gimbal bearings or, better, on a hardened steel or jewel cone and cup bearing. The skirt of the aluminum cylinder limits the sideways tipping to about 20°, but this has not been seriously restrictive. By moving both large rings up or down, the center of gravity of the unit can be placed at the center of support or at a distance from it along the main axis, as desired. In addition, there is provision for a semipermanent transverse adjustment of the center of support. This "top" is given sufficient spin by merely twirling it with the fingers, *e.g.*, 5 to 10 rps.

By means of the top and a stroboscope, many features of standard gyroscopic theory are easily checked. Various precession rates, including zero, can be obtained. Nutations are sustained, undamped (for the rigid body), over long periods so that it is easy to see, for example, that nutation frequency is higher than spin for $A/B > 1$ and lower for $A/B < 1$. It was quickly demonstrated that an effective nutation damper was, in fact, an undamper when $A/B < 1$. It is noted, however, that if the friction in the spin bearing becomes too high it has a damping effect on nutations, and test results under this condition will be deceiving.

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[1] R. G. McCarty, "Designing small space loads," *Astronautics*, vol. 4, p. 24; May, 1959.

A sample of the damping devices tested with this top is shown in Fig. 2. It has been found quite easy to explore or test out ideas for new types of damping devices. Qualitative tests of an idea are usually sufficient to answer most immediate questions. By means of a stop watch and rough observations of nutation amplitude semiquantitative results can be obtained. For quantitative determinations motion pictures can be taken. The frame rate need not be high, 64 to 90 frames per second is sufficient, but a very narrow shutter angle such as 10° should be employed to minimize blurring. Although it may be somewhat tedious, measurements can be made of 1) spin angle, 2) attitude of the axis of the top, 3) position of the damping device or its components. These data can be plotted or analyzed to show quantitatively the relations between spin, nutation, precession, and functioning of damper parts.

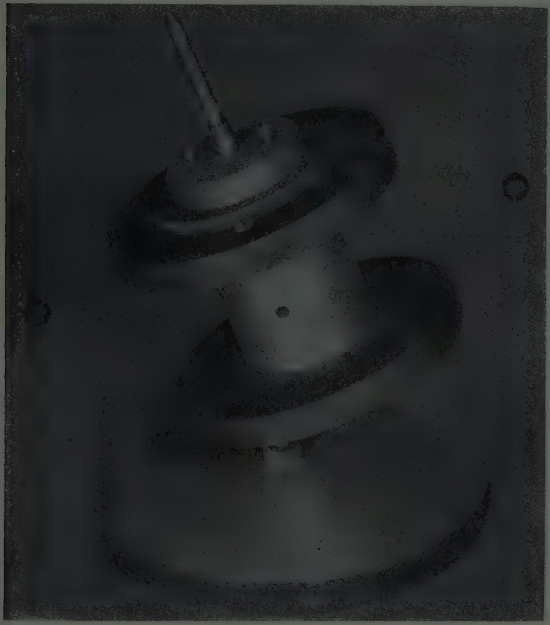

Fig. 1—Space vehicle dynamical simulator.

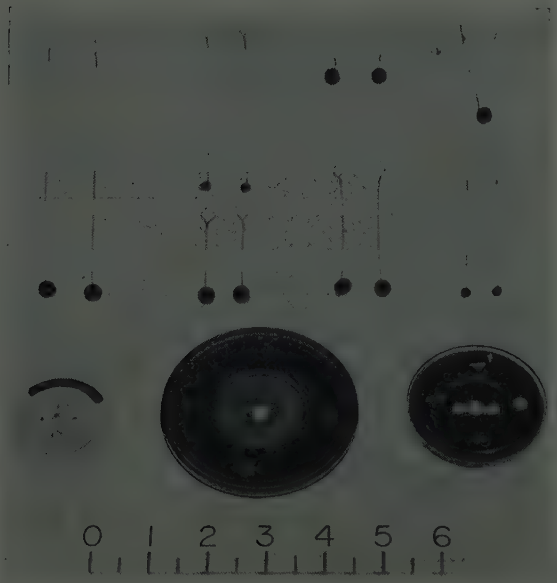

Fig. 2—Experimental nutation damper.

ANALYSIS

Although photographs have been taken, experiments to date have only been semiquantitative. They have, however, indicated that there is more than one mechanism, some probably more effective at large amplitudes, and others at small. Studies of various aspects of the problem have been made[2–9] which in general tend to support this observation. Theories which depend entirely on frictional dissipation to absorb nutational energy do not seem to be nearly fast enough.

A more recent analysis[9] has probably followed the most rigorous approach, and in it the effect of some of the approximations finally made were tested by analog computor methods. The specific problem treated is that of a single small mass constrained to move in a plane normal to the axis of the top and at a fixed radius from this axis. This can be, for example, a small mass on the end of a light, rigid arm hung on the axis of the top, or a small drop of mercury in a channel concentric with the axis.

The moving coordinates chosen are aligned with the symmetry axis of the rotating rigid body, *i.e.*, the top, with the origin at the center of gravity of the top. However, the coordinate system rotates with the small damper mass instead of with the top (Fig. 3). Euler rotations from the nonrotating coordinates, x_0, y_0, z_0, are shown in Fig. 4 as follows: 1) rotation, ϕ, about the fixed vertical axis, z_0; 2) rotation, θ, about the new x, which will be the line of nodes; 3) rotation, ψ, about the

[2] F. H. Davis, "Gyro Nutation Damping," U. S. Naval Ord. Test Sta., China Lake, Calif., NOTS Internal Memo.; March 10, 1959.

[3] F. H. Davis, "Oscillations of a Disc-Shaped Pendulum Mounted on a Spinning Body," U. S. Naval Ord. Test Sta., China Lake, Calif., NOTS Internal Memo.; July, 1959.

[4] W. F. Cartwright, U. S. Naval Ord. Test Station, China Lake, Calif., unpublished NOTS Internal Memos. on the Nutation Damper; July 12, 1958, July 22, 1958, and others.

[5] E. E. Rogers, "A Mathematical Model for Predicting the Damping Time of a Mercury Damper," U. S. Naval Ord. Test Sta., China Lake, Calif., NOTS Internal Memo. No. IDP 565; March 2, 1959.

[6] D. J. Stewart, "Nutation Damper," U. S. Naval Ord. Test Sta., China Lake, Calif., U. S. Patent No. 2,734,384; February 14, 1956.

[7] J. M. Boyle, "Invention Disclosure for Nutation Damper," U. S. Naval Ord. Test Sta., China Lake, Calif., Navy Case No. 25,818; September 10, 1957. (Currently inactivated.)

[8] G. F. Carrier, "The Theory of the Annular Damper," Space Technology Labs., Los Angeles, Calif., Rept. No. GM-TM-0165-00265; July 21, 1958.

[9] A. V. Pratt, "Theory of the Unsymmetrical Nutation Damper," U. S. Naval Ord. Test Sta., China Lake, Calif., NOTS Internal Memo. No. TN-4065-63. (In press.)

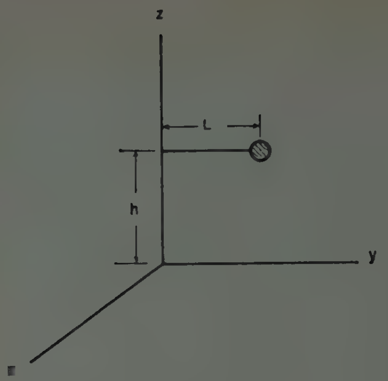

Fig. 3—Damper mass, m, fixed in rotating coordinate system.

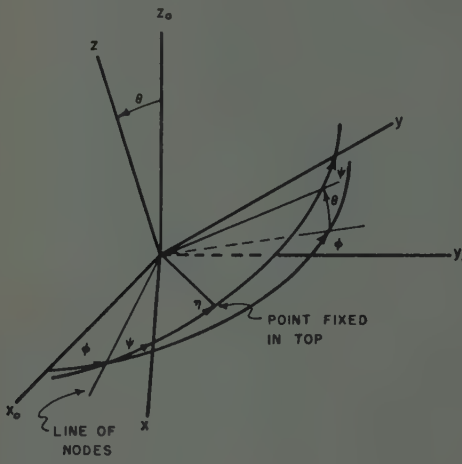

Fig. 4—Fixed and moving coordinates showing Euler angles and η.

final z, which is also the axis of symmetry of the top. Finally, to reach a fixed point in the top from the x, z plane, an additional rotation, η, about z is required.

It is assumed that no external forces or moments are acting, and the frictional coupling between the small mass and the top produces a force *not* resulting from a potential function. Therefore the Lagrangian of the system is simply the sum of the kinetic energies of the top and the small mass. This is

$$T = \tfrac{1}{2}[A\Omega^2 + B(\omega_x^2 + \omega_y^2) + MV_M^2 + mV_m^2],$$

where ω_x, ω_y, (and ω_z, used later) are the components of the total angular velocity of the moving coordinate system along the moving axes indicated by the subscripts; Ω is the component of angular velocity of the top along z, *i.e.*, spin (note that $\Omega=\omega_z+\dot{\eta}$); V_M, V_m are velocities of the top and of the damper, respectively; M, m are masses of the top and of the damper respectively; and A, B are polar and transverse moments of inertia, respectively, of the top. This reduces to

$$T = \tfrac{1}{2}[A\Omega^2 + B(\omega_x^2 + \omega_y^2) + \bar{m}(\Omega \times r_m)^2],$$

where $\bar{m}=m\ [M/(M+m)]$ is a "reduced" damper mass, and r_m is the distance from the center of gravity (cg) of the top to the damper mass.

The generalized coordinates, q_i, of the Lagrange equations are the three Euler angles of the moving coordinate system, in which the damper mass is fixed, and the angle η between x and a reference point in the x, y, plane in the top. In terms of these the four angular velocities in the expression for T are

$$\omega_x = \dot{\phi}\sin\theta\sin\psi + \dot{\theta}\cos\psi$$

$$\omega_y = \dot{\phi}\sin\theta\cos\psi - \dot{\theta}\sin\psi$$

$$\omega_z = \dot{\phi}\cos\theta + \dot{\psi}$$

$$\Omega = \omega_z + \dot{\eta}$$

in which dots show differention with respect to time. These are substituted into T and the result differentiated as indicated in the modified Lagrange equations:

$$\frac{d}{dt}\left(\frac{\partial T}{\partial \dot{q}_i}\right) = M_{q_i}.$$

The M_{q_i} are all zero except that

$$M_\eta = -k\dot{\eta},$$

in which simple viscous friction alone (with coefficient, k) is assumed. The four equations which result can be reduced to the following in which it has been possible to eliminate the trigonometrical functions.

$$\dot{\Omega} = -\frac{k}{A}(\Omega - \omega_z).$$

$$\dot{\omega}_z = \frac{\bar{B}k}{\bar{B}C - D^2}(\Omega - \omega_z) + \frac{AD}{\bar{B}C - D^2}\Omega\omega_x + \omega_x\omega_y.$$

$$\dot{\omega}_y = -\omega_z\omega_x + \frac{AC}{\bar{B}C - D^2}\Omega\omega_x + \frac{Dk}{\bar{B}C - D^2}(\Omega - \omega_z).$$

$$\dot{\omega}_x = \frac{\bar{B} - C}{\bar{B} + C}\omega_z\omega_y - \frac{A}{\bar{B} + C}\Omega\omega_y - \frac{D}{\bar{B} + C}(\omega_z^2 - \omega_y^2).$$

Here the new symbols are as follows:

$$\bar{B} = B + \bar{m}L^2$$

$$C = \bar{m}h^2$$

$$D = \bar{m}hL.$$

h, L are the y and z coordinates, respectively, of the damper mass.

These equations are exact for this particular problem, without small angle or other limiting approximations. The case should be good for a vehicle in space without (gimbal) bearing friction if the frictional coupling to the damper mass is indeed entirely through viscous friction.

If the nutation angle, θ, is kept small and if the damper mass is much less than the mass of the top, the following simplified equations are obtained:

$$\dot{\Omega} = 0.$$

$$\dot{\omega}_z = -\frac{k}{C}(\omega_z - \Omega) + \frac{AD}{BC}\Omega\omega_x.$$

$$\dot{\omega}_x = \omega_z\omega_y - \frac{A}{B}\,\Omega\omega_y - \frac{D}{B}\,\omega_z^2.$$

$$\dot{\omega}_y = -\,\omega_z\omega_x + \frac{A}{B}\,\Omega\omega_x.$$

The difference between these two sets of equations in determining angular velocities was shown to be small under the limiting conditions by analog computor simulations.

In studying the equations for nutation damping one can examine the angular velocities to learn, for example, how fast ω_x and ω_y become small under various initial conditions and relationships of the parameters. However, these angular velocities are difficult to measure accurately in an experiment, and an expression for the nutation angle itself is greatly to be desired. After much algebra but without further limiting assumptions the following differential equations were obtained.

$$\ddot{R} = -\frac{k}{c}(\dot{R} - \Omega) + \frac{AD}{2BC}\,\Omega(\dot{Z}e^{-iR} + \dot{\bar{Z}}e^{iR}),$$

$$\ddot{Z} - i\,\frac{A}{B}\,\Omega\dot{Z} = -\,\frac{D}{B}\,\dot{R}^2e^{iR},$$

in which $R \equiv \phi + \psi$ is approximately the rotation angle of the coordinate system (and the damper mass). Note that

$$\dot{R} \doteq \omega_z$$

$$Z \equiv \sin\theta(\cos\psi + i\sin\psi)$$

so that

$$|Z| = \sin\theta \doteq \theta.$$

It will be remembered that, in the approximation, the spin of the top, Ω, is considered to be constant.

It is evident that these equations are extremely difficult to solve in the general sense. However, three different simplified probable cases were examined. The first is the observed stable situation in which the damper does not rotate relative to the top. The expected limit cycle results in which $|Z| \doteq \theta_0 =$ constant.

For the second condition the damper oscillates at a small amplitude with respect to a fixed point in the spinning top. This is called the "slow damping" case. The solution consists of a constant term and successive terms decreasing the nutation angle with time. This can be put into exponential form and the value for the time constant (in which the amplitude is reduced to $1/e$ of an initial value) is given by

$$t_C = \frac{2(A - B)BC^2\left[\left(\frac{k}{C}\right)^2 + \left(\frac{A - B}{C}\right)^2\Omega^2\right]}{AD^2\Omega k}.$$

Analog computer solutions indicate that the time given by this formula is a little too long.

A third solution attempted, called the "fast damping" case postulates that the damper mass rotates at the nutation frequency. No reasonable exponential form for the solution of this case could be written. However, the nutation amplitude during any single nutation cycle is indicated approximately by

$$\theta^2 = \theta_0^2 - \left(\frac{2k(A - B)}{AB} - \frac{2\pi D^2A\Omega}{B^3}\right)t.$$

This relation appears to hold fairly well until θ approaches a minimum value below which the solution no longer exists.

Investigation is being continued with this simplified model which should further the accurate analysis of more practical designs. The attack is threefold, employing the experimental model and the analog computer to substantiate and to stimulate mathematical analysis. It appears to date that in the most powerful mechanisms the damper plays the part of a catalyst, an engine or a coupling which causes or makes possible the transfer of energy or momentum from the nutational vibration to the axial spin.

INTUITIVE DESIGN CRITERIA

The simple device discussed analytically above, which may consist of a small weight hung on the shaft of a top by a length of fine, stiff wire, provides good nutation damping (or undamping for $A/B < 1$). But a small unbalance is created causing the stable rotation to be about some axis other than the shaft axis and at random orientation from it. This makes it impossible to arrange internal components of a space vehicle so that any one part will always be on the final stable axis of rotation. However, the orientation can be predetermined if a fixed unbalance is introduced in the top or, similarly, if the pivot point for the pendulum-like damper is a pin offset to one side of the top's center shaft. This suggests the Frahm vibration absorber as discussed by DenHartog.[10] To bring the stable axis of rotation back to the shaft, *i.e.*, the axis of symmetry, a counterbalance weight can be introduced opposite to the pendulum as proposed by Davis.[11] The obvious extension of this is to employ two diametrically opposed pivot pins and pendulums designed to counterbalance each other when in the stable state of rotation of the top. As described later, the stable state of rotation is controlled by more design parameters than just the weight of these opposed pendulums. However, it is sufficient now to note that, for certain conditions this double pendulum damper approaches the effectiveness of the mercury damper in tests made with this top. Even though the complete and rigorous theory is not available for these dampers, it may be of interest to show here a naive or intuitive approach to the design of both the pendulum and the mercury dampers which appears to work. In both cases the treatment is only slightly modified from that indicated by Davis.

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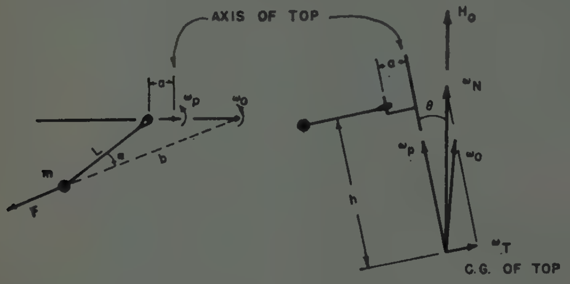
[10] J. P. DenHartog, "Mechanical Vibrations," McGraw-Hill Book Co., Inc., New York, N. Y., 3rd ed.; 1953.

[11] This and the offset pivot were suggested by F. H. Davis of the U. S. Naval Ord. Test Sta., China Lake, Calif.

Pendulum Nutation Damper

First considered is the pendulum type of damper. The pendulum is in a centrifugal force field with lines radiating from the instantaneous center of rotation, ω_0. In Fig. 5(a) this is shown as the force F, on mass, m, directed along a radius, b, from ω_0. The axis of the top is labelled with its spin, ω_p, and the peg is shown at distance, a, from ω_p. We assume for simplicity that these points remain in a line, which they actually do *only* if the nutation angle, θ, is constant and there is no damper attached. This line is in the plane containing the angular momentum and spin vectors as shown in Fig. 5(b).



(a) (b)

Fig. 5—Pendulum damper parameters.

With this simplified picture, it is easy to find the following formula for the frequency of the pendulum which should be considered only for very small θ:

$$\text{frequency} \equiv \frac{\omega_\beta}{2\pi} = \frac{\omega_0}{2\pi}\sqrt{\frac{a}{L} + \frac{hA}{LB}\tan\theta}.$$

Here we have employed the classical relation between the nutation rate, ω_N, its angle, θ, and the polar and transverse moments of inertia, A and B, respectively, *i.e.*,

$$\frac{\omega_N}{\omega_p} = \frac{A}{B\cos\theta}$$

and the relation illustrated, *i.e.*,

$$\omega_T = \omega_N \sin\theta.$$

We reason now that if this is to be a vibration absorber, the pendulum frequency should be tuned for the disturbing frequency, namely the nutation frequency, ω_N. However, since the pendulum is spinning with the top it will not "see" ω_N but the difference between this and the spin frequency. Therefore, we set

$$\omega_\beta = \omega_N - \omega_p,$$

and solve for the relation between the parameters to find that the length of the pendulum should be

$$L \doteq \frac{a + Ih\theta}{(I-1)^2} \rightarrow \frac{a}{(I-1)^2}$$

where $I = A/B$ is the ratio of the polar to the transverse moments of inertia and where small angle approximations have been made for functions of θ; also, we have let $|\omega_0| = |\omega_p|$. The formal or mechanical substitution of terms shows that the tuning depends on the amplitude of nutation, θ. However, since the relation is suspect for significant values of θ, damper design should probably be based on the first term only, as indicated by the arrow.

If I is not far above unity, the required L may be excessively long. A compound pendulum having radius of gyration, ρ, about its center of gravity can then be designed according to the following:

$$L = \frac{a + Ih\theta}{(I-1)^2}\,\frac{1}{(1 + \rho^2/L^2)}.$$

This pendulum can be a disk or a ring attached so that ρ/L is large.

As has been noted, an arrangement in which two of these pendulums are hung on diametrically opposed pegs forms an extremely effective balanced damper. The two pendulums can be tuned to different frequencies to broaden the most effective damping range; but the centrifugal force created by each, in the stable state, should be the same for good balance.

Mercury Nutation Damper

For the mercury damper the approach is similar. As noted earlier, this damper consists of one or more circular channels affixed concentrically about the axis of the top. For a space vehicle they would be mounted some distance from the center of gravity. Each channel is filled with a certain amount of mercury as discussed in the following. There must be sufficient mercury to form a complete ring in the channel when the top is spinning smoothly about its axis.

Let the radial depth of mercury in this ring be d and call the centrifugal acceleration g^*. Then the velocity of a wave in the mercury is

$$V = \sqrt{g^* d}.$$

This holds when the depth is small compared to the wavelength; the velocity will be lower if the depth becomes proportionately greater. If this velocity is divided by the circumference of the mercury surface we obtain the frequency with which a wave travels around the channel circle. The principle is to match this frequency to the disturbance frequency "seen" by the mercury; that is, $(\omega_N - \omega_p)$ as with the pendulum. Here $g^* = L\omega_0^2$, so that

$$\omega_M = \frac{V}{L} = \frac{\sqrt{L\omega_0^2 d}}{L} = (\omega_N - \omega_p).$$

Then $d=L(I-1)^2$ is radial depth of liquid. The symbols are as before except that L represents the radius from the top axis to the inner mercury surface when the top is "sleeping." The same small angle approximations have been made as for the pendulum. Two or more channels, if used, can be tuned to somewhat different frequencies to broaden the effective range.

An alternative damping mechanism for the mercury is also suggested by Davis.[2] Since the nutation rate is greater than the spin, the mercury is driven to rotate in its channel faster than the top. By the frictional coupling the mercury tends to increase the spin of the top. This mechanism would transfer nutational energy into spin energy, damping nutational amplitude.

A DESIGN TEST

A very interesting criterion has been found recently for the nature of the stable conditions of the top with two diametrically opposed pendulums; and, as will be seen, for space vehicles a certain combination of design parameters is definitely to be avoided.

It was first observed that the top, when fitted with a particularly long but well balanced pair of pendulums, damped quickly to a constant or stable amplitude (apparently) of nutation. However, when this performance was viewed under the stroboscopic light it was found that the two pendulum arms had finally assumed an attitude in which the angle between them was not 180° (Fig. 6). The resulting unbalance caused the assembly to rotate, stably, about an axis at an angle to the shaft of the top. The true nutation angle had damped to zero but there was a new principal axis, not the shaft of the top, about which final rotation occurred.

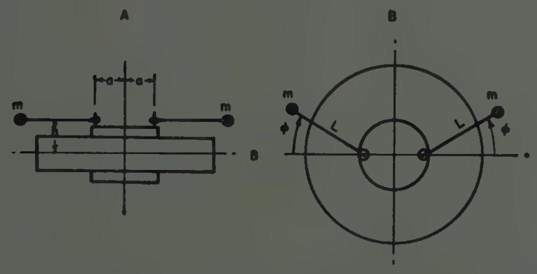


Fig. 6—Twin pendulums at angle ϕ.

This phenomenon was studied and the results are reported by Haseltine.[12] Instead of a detailed attack on the motion of the coupled rigid bodies, general energy arguments are used. It is demonstrated that there exists a minimum energy state and that this occurs when the pendulums are in a fixed attitude relative to the top. Under the conditions for this fixed attitude, it is shown that the assembly must rotate about the axis of greatest moment of inertia. Further, under this condition and for certain combinations of the design parameters, the pendulums will be deflected toward each other as actually observed with the top.

An approximate expression is derived for $\cos\phi$ where ϕ is the angle that each pendulum deviates from the diameter (extended) through the supporting pegs (Fig. 6). It is

$$\cos\phi = \text{the lesser of 1 or}\left(\frac{a(A-B)}{2mh^2L}\right),$$

where

a = distance from the top axis to the pendulum supporting pegs,
m = mass of one pendulum bob,
h = distance from center of gravity of the top to the plane of the pendulum supports,
L = length of the pendulum, and
A, B = respectively, the polar and transverse moments of inertia of the top.

It has been noted that if we know by experiment, or otherwise, that this stable condition of off-axis weights can exist, a simple derivation of the formula for the angles, ϕ, is possible. In Fig. 7 note that the two pendulums fall on lines crossing at the stable axis of rotation, A', as they must, since centrifugal force holds them in this position. The center of gravity (cg) of the two bobs falls at the point noted by $2m$ which is the distance h along the axis A, from the cg of the top and a distance ($L\sin\phi$) away from that axis. The rotation, θ, of the principal axis of inertia as the result of this mass is, approximately

$$\tan 2\theta = 2\,\frac{(\text{mass} \times d_1 \times d_2)}{(A-B)} = 2\,\frac{2mhL\sin\phi}{(A-B)}.$$

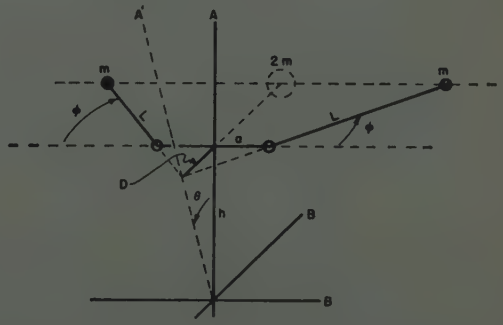


Fig. 7—Twin pendulums, cos ϕ derivation.

If θ is small, the factor 2 can be cancelled.

Multiplying $\tan\theta$ by h gives the distance, D, in Fig. 7, which can also be obtained by

$$D = a\tan\phi.$$

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[12] W. R. Haseltine, "Passive Damping of Wobbling Satellites," U. S. Naval Ord. Test Sta., China Lake, Calif., NOTS 2306 NAVORD 6579. (In press.)

The following can now be written:

$$D = h \tan \theta = a \tan \phi = \frac{2mh^2 L \sin \phi}{(A - B)}$$

or cancelling $\sin \phi$ from the last two and solving for $\cos \phi$,

$$\cos \phi = \frac{a(A - B)}{2mh^2 L},$$

as before.

It is quite evident that vehicle design should keep this expression for $\cos \phi$ comfortably greater than unity. For, even though we might devise a way to insure that the pendulums always go to the same side of the axis, A, so that we would know where A' will fall and design accordingly, the angle ϕ would be sensitive to changes in the design parameters and probably would not be very stable.

We therefore have a criterion for conditions to avoid in designing rotating space vehicles with pendulum dampers. Also, this development may serve as a warning that a similar situation might arise for other nutation damping schemes. It has, in fact, been shown that a similar condition can occur when four equal weights are used instead of two. The condition for good behavior is then similar but slightly more stringent.

Measurement of the Doppler Shift of Radio Transmissions from Satellites*

GEORGE C. WEIFFENBACH†

Summary—**This paper describes a receiving station network that was designed to produce Doppler data for use in a satellite Doppler navigation system. As a result of practical exigencies, this equipment is not optimum for its intended purpose, and a continuous effort is being maintained to improve the system. Current and projected changes are described briefly. Some experimental results obtained with the present system are presented.**

I. Introduction

THE function of the system which is described herein is to measure the Doppler shift of radio signals broadcast from a satellite. In particular, this system must detect the satellite transmissions, measure their frequency as a function of time, and reduce these measurements to a form suitable for transmission to a central computer. A previous paper[1] outlines the over-all navigation system of which the receiving station complex is one part.

The configuration of the receiving stations used for tracking is governed by the navigation system requirement for real time predictions of satellite positions to very high accuracies. Present ignorance of the geoid plus the accumulation of errors in digital computers will prevent, initially, the extrapolation of even an exact present orbit for more than a day or two into the future without a significant loss of accuracy. (Note that this is independent of the particular kind of observational data that was used to obtain the "present" orbit.) This fact necessitates the acquisition of dense sets of accurate data; *i.e.*, in a time which is substantially less than the "extrapolation time" of one day, one must accumulate observational data of sufficient quantity and quality that a very accurate present orbit may be computed. Furthermore, if a true real time *prediction* of satellite position is to be made (the only kind that is of interest in an operational navigation system), this data must be reduced and transmitted to the computing center in a time which is short with respect to the "extrapolation time." For these reasons it was decided that the data gathering or receiving station network must necessarily be capable of accurate and automatic reduction of the Doppler shift information to digital form. It must be automatic so that large quantities of data can be handled quickly and without serious error; digital so that it can be transmitted rapidly and without significant degradation.

The configuration of the navigation station depends on the accuracy required, with optimum performance necessitating instrumentation roughly equivalent to that used in the tracking stations.

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[1] W. H. Guier and G. C. Weiffenbach, "A satellite Doppler navigation system," this issue, p. 507.

II. Description of Tracking Station

Fig. 1 is a flow diagram of the apparatus that has been assembled to meet the above requirements. During the pass of an appropriate satellite, one can feed the signal from an antenna into one end of the equipment and simultaneously extract from the other end a strip of TWX tape containing an accurate description of the Doppler shift of the satellite transmission as a function of time.

Fig. 2 illustrates the receiving section, which consists of an antenna and a commercial receiver (Nems-Clarke Model 2501) and a phase-locked tracking filter (Interstate Electronics Model IVA). The function of this section is to pick up the weak satellite signals, amplify them, separate them from noise, and subtract from them an accurately known frequency (the "reference" or "injection" frequency) so that they form an appropriate input to the data reduction unit. It has been determined that the present equipment can perform this function satisfactorily with an input signal somewhat less than 0.02 microvolt, and indeed, using a simple vertical half wave antenna, successful operation has been obtained in observing the 10-mw 108-mc transmitter of 59 η whenever it was above the horizon.

Fig. 3 illustrates the time and frequency reference unit, which generates a local time base and a set of standard frequencies. The time is used to indicate when a particular Doppler measurement was made. The standard frequencies are used for the receiver injection frequency, for a tape recorder reference, and for the "meter frequency" which is used to measure the Doppler period in the data reduction unit. It is clearly necessary that a common time and frequency base be used throughout the navigational system. The most convenient method of accomplishing this is to synchronize the local references to the standard time and frequency broadcasts of the National Bureau of Standards WWV. (As pointed out previously,[1] the navigational satellite itself could be used to "transport" WWV or any other suitable time standard to any point on the globe, by transmitting time pulses plus correction data from the satellite at intervals of perhaps one minute.) The "missing pulse" detector is used to start the local clocks at the first second marker after the missing pulse in the WWV transmission. Corrections are made thereafter by adding or subtracting 1 mc cycles (μsec) with the motor-driven phase shifter. The continuously changing propagation paths of WWV transmissions preclude direct frequency comparisons of sufficient accuracy. The long term drift is monitored by observing deviations of the time base relative to WWV over a period which is long enough to average out all effects of propagation variations. A second frequency standard is maintained at each station to provide a short-term (seconds to minutes) monitor of local frequencies.

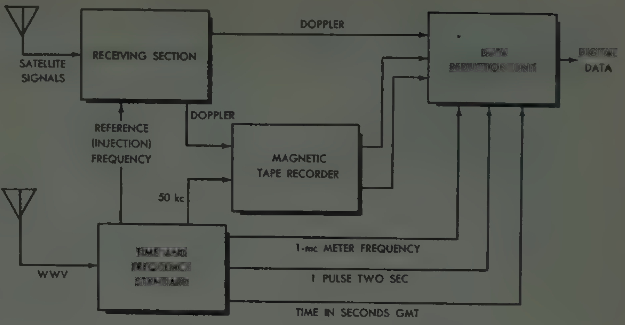
SATELLITE SIGNALS
RECEIVING SECTION
DOPPLER
DATA REDUCTION UNIT
DIGITAL DATA
DOPPLER
REFERENCE (INJECTION) FREQUENCY
MAGNETIC TAPE RECORDER
50 kc
WWV
TIME AND FREQUENCY STANDARD
1-mc METER FREQUENCY
1 PULSE TWO SEC
TIME IN SECONDS GMT

Fig. 1—Over-all flow diagram for receiving stations.

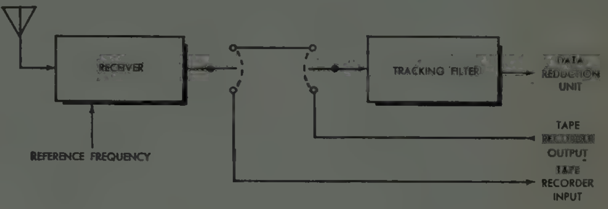
RECEIVER
TRACKING FILTER
DATA REDUCTION UNIT
REFERENCE FREQUENCY
TAPE RECORDER OUTPUT
TAPE RECORDER INPUT

Fig. 2—Simplified flow diagram for receiving section.

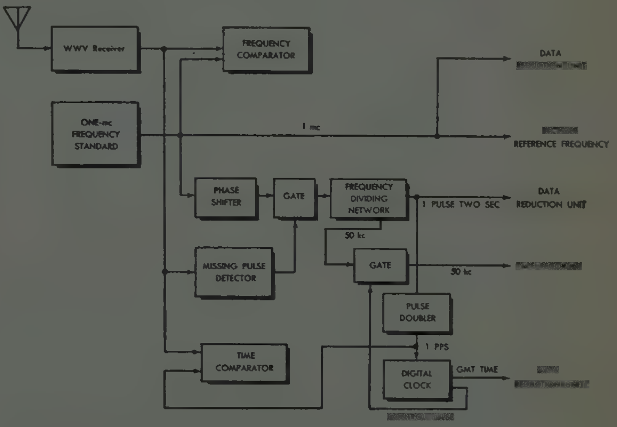
WWV Receiver
FREQUENCY COMPARATOR
DATA REDUCTION UNIT
ONE-mc FREQUENCY STANDARD
1 mc
REFERENCE FREQUENCY
PHASE SHIFTER
GATE
FREQUENCY DIVIDING NETWORK
1 PULSE TWO SEC
DATA REDUCTION UNIT
50 kc
MISSING PULSE DETECTOR
GATE
50 kc
PULSE DOUBLER
1 PPS
TIME COMPARATOR
DIGITAL CLOCK
GMT TIME

Fig. 3—Simplified diagram of time and frequency reference unit.

The accuracies attained in the time and frequency units directly determine (together with satellite transmitter stability, etc.) the accuracy of the basic Doppler data. Table I lists the present estimates of the accuracies and stabilities needed to achieve a navigational accuracy of $\frac{1}{2}$ mile or better. All of the stipulated requirements are met in the present system using commercial components.

Fig. 4 illustrates the data reduction unit. Its function is to measure the Doppler frequency (actually it measures period), correlate it with time and punch the result on TWX paper tape. This is accomplished by the following sequence of events:

1) Timing pulse (1 pulse every 2 seconds) opens gate to preset counter.
2) Preset counter puts out gate-opening pulse at instant of first positive-going zero crossing of Doppler signal.

TABLE I

Reference frequency	Stability:	10^{-9}/seconds 10^{-9}/15 minutes
	Accuracy:	2×10^{-9}
Meter frequency	Stability:	10^{-7}/seconds 10^{-7}/15 minutes
	Accuracy:	2×10^{-7}
Clock	Stability:	10^{-8}/15 minutes
	Accuracy:	5 ms
Tape Recorder Reference (50 kc)	Stability:	10^{-7}/seconds 10^{-7}/15 minutes
	Accuracy:	2×10^{-7}

Time and frequency accuracy and stability requirements (based on assumption that WWV is exact). The U. S. Naval Observatory, and others, provide corrections to WWV broadcasts which are used in the navigation system central computing facility.

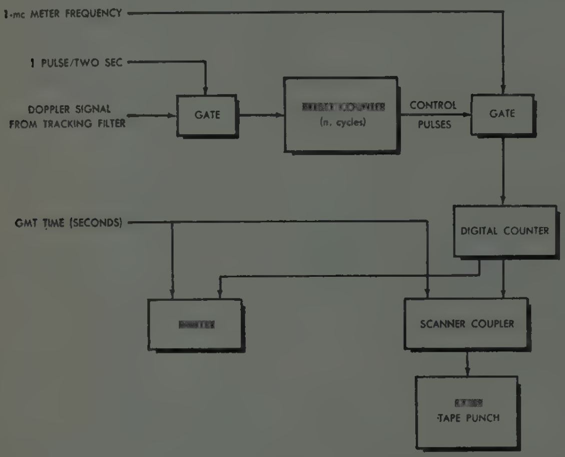

Fig. 4—Simplified diagram of data reduction unit.

3) Preset counter puts out gate-closing pulse at instant of n_cth positive-going zero crossing of Doppler signal.
4) Gate controlled by preset counter is open for duration T of n_c Doppler cycles. T is measured by observing the number of 1-mc cycles (μsec) that pass through gate. This number is registered on digital counter.
5) The scanner coupler then reads the time on the digital clock (in the time and frequency reference unit) and the meter frequency counter reading, and feeds these numbers serially to the paper tape punch.
6) The paper tape punch punches the appropriate set of numbers on paper tape.

The output of this section is an eleven digit number in which the first five digits indicate time in seconds after midnight GMT, and the last six digits indicate the period (in μsec) of the first n_c Doppler cycles following this time reading. The number n_c is selected so that T for the lowest expected value of the Doppler signal is less than, but nearly, one second. The scanning and readout take about 600 ms. One data point is obtained every two seconds. The only significant error introduced in this digitalizing process is the ±1 count ($=\pm1$ μsec) resolution of the meter frequency counter, which results in a frequency error of $\pm(f^2/n_c)\times10^{-6}$ cps, where f is the (audio) Doppler input to the preset counter.

Fig. 5 illustrates the magnetic tape recorder unit. This section provides storage for Doppler data pending availability of the data reduction unit and tracking filter, so that a receiving station with only one each of the latter can obtain data for a multifrequency satellite. The Doppler signals directly out of the receivers are recorded simultaneously with a 50-kc reference. A time reference is obtained by gating the 50 kc, *i.e.*, starting the 50 kc on the recording by means of a pulse from the real time clock. By counting down the 50 kc to 1 pulse per second and feeding this into a prefilled (playback) clock the latter will operate in synchronism with the original clock readings at the time the Doppler was recorded. By multiplying the 50 kc up to 1 mc for use as the meter frequency, the effects of variable tape speed, from record through playback, should be removed.

One additional unit is planned but is not in operation at the present time. This is the analog refraction correction unit shown in Fig. 6. This unit makes use of the known dispersion properties of ionospheric refraction, *i.e.*, that the effective index of refraction of the ionosphere, n, is given to first approximation by the formula[1]

$$n = \sqrt{1 - f_N^2/f_i^2}$$

where

f_N = plasma frequency = $\sqrt{Ne^2/\pi m}$
N = electron density in electrons/cc
e = electronic charge in esu
m = electron mass in grams
f_i = transmitter frequency.

Using this relationship, the observed Doppler shift can be approximated by a power series in $1/f_i$ as[1]

$$\Delta f^{(i)}(t) = \Delta f_0^{(i)}(t) + \Delta f_1^{(i)}(t) \qquad (1)$$

where

$\Delta f^{(i)}(t)$ = Doppler shift of f_i as received,
$\Delta f_0^{(i)}(t)$ = "vacuum" Doppler shift, *i.e.*, as it would be in absence of ionosphere,
$\Delta f_0^{(i)}(t) = f_i\Delta f_0^{(0)}(t) = f_i\dot{\rho}(t)/c$,
$\dot{\rho}(t)$ = time rate of change of slant range from ground station to satellite,
c = vacuum velocity of light,
$\Delta f_1^{(i)}(t)$ = first order "error term" resulting from ionospheric effects,
$\Delta f_1^{(i)}(t) = \alpha(t)/f_i$
$\alpha(t)$ = power series coefficient for $1/f_i$ term.

Since $\Delta f_0^{(0)}(t)$ and $\alpha(t)$ are independent of frequency, (1) can be used in conjunction with measurements of $\Delta f^{(i)}(t)$ at two frequencies to construct two simultaneous equations that can be solved for $\Delta f_0^{(0)}(t)$ or $\alpha(t)$ by

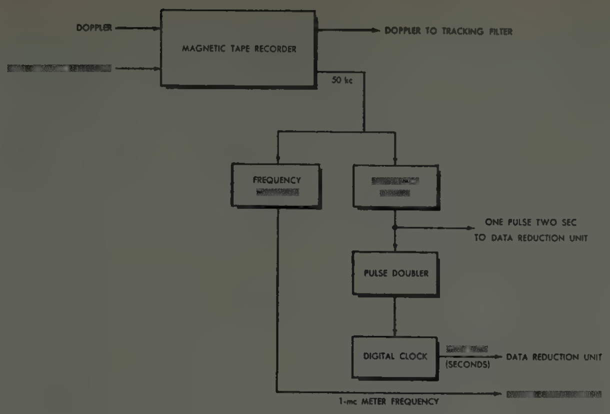
DOPPLER
MAGNETIC TAPE RECORDER
DOPPLER TO TRACKING FILTER
50 kc
FREQUENCY
ONE PULSE TWO SEC
TO DATA REDUCTION UNIT
PULSE DOUBLER
DIGITAL CLOCK
(SECONDS)
DATA REDUCTION UNIT
1-mc METER FREQUENCY

Fig. 5—Simplified diagram of magnetic tape recorder unit.

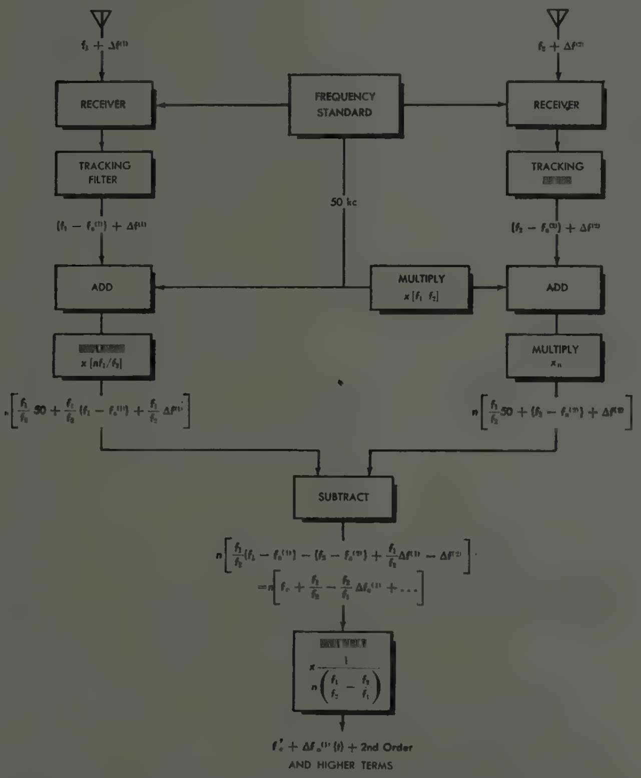
$f_1 + \Delta f^{(1)}$
$f_2 + \Delta f^{(2)}$
RECEIVER
FREQUENCY STANDARD
RECEIVER
TRACKING FILTER
TRACKING
50 kc
$(f_1 - f_c^{(1)}) + \Delta f^{(1)}$
$(f_2 - f_c^{(2)}) + \Delta f^{(2)}$
ADD
MULTIPLY $\times [f_1\ f_2]$
ADD
$\times [nf_1/f_2]$
MULTIPLY $\times_n$
$n\left[\frac{f_1}{f_2} 50 + \frac{f_1}{f_2}(f_1 - f_c^{(1)}) + \frac{f_1}{f_2}\Delta f^{(1)}\right]$
$n\left[\frac{f_1}{f_2} 50 + (f_2 - f_c^{(2)}) + \Delta f^{(2)}\right]$
SUBTRACT
$n\left[\frac{f_1}{f_2}(f_1 - f_c^{(1)}) - (f_2 - f_c^{(2)}) + \frac{f_1}{f_2}\Delta f^{(1)} - \Delta f^{(2)}\right]$
$= n\left[f_c + \frac{f_1}{f_2} - \frac{f_2}{f_1}\Delta f_0^{(1)} + \ldots\right]$
$\times \frac{1}{n\left(\frac{f_1}{f_2} - \frac{f_2}{f_1}\right)}$
$f_c' + \Delta f_0^{(1)}(t)$ + 2nd Order
AND HIGHER TERMS

Fig. 6—Analog instrumentation for extracting $f_0^{(1)}(t)$, the "unrefracted" Doppler shift for the satellite transmitter frequency f_1. The two satellite transmitter frequencies, f_1 and f_2, should be coherent. (Note: $f_{c'}$ is a known constant. The final multiplication is done in the central computer.)

elimination. The instrumentation shown in Fig. 6 will perform this elimination to produce $\Delta f_0^{(i)}(t)$, the "vacuum" Doppler. If the approximations are sufficiently accurate, this will provide refraction-free data for the computer. Note that when f_1 is the second harmonic of f_2, for example, all of the operations shown in the figure can be performed with standard circuits.

The instrumentation described in the preceding section constitutes a tracking station for the present satellite navigation system. Clearly, any one of these stations can function as a navigating station simply by using its output data for computing a navigational fix instead of sending the data to the central computer for use in orbit determination.

III. PRESENT TRACKING STATION SYSTEM

At the time of this writing, five receiving stations, each as described in Section II, have been assembled to form a system whose function is to gather the data for precision tracking of a navigational satellite. These stations are located in Howard County, Maryland; Argentia, Newfoundland; Seattle, Washington; the Defense Research Laboratory of the University of Texas, Austin, Texas; and the Physical Sciences Laboratory of the New Mexico State University, Las Cruces, New Mexico. A sixth installation which is only instrumented to produce magnetic tape for reduction at one of the complete stations is currently operated by the Radio Department of the Royal Aircraft Establishment at Lasham, England.

It is hoped that it will be possible to enlarge this network to provide more extensive geographic coverage, which is necessary to realize the full potential of the system to produce accurate ephemerides. The present locations were selected to provide a "hard core" of stations which would not get entangled in acute logistic and communications problems, but would, nonetheless, cover a reasonably wide area. It has been possible to transmit all data by commercial TWX, and only at Argentia has it been necessary to use a leased line.

IV. COMMENTS

The equipment described above has functioned satisfactorily in that it has produced data in actual practice that is sufficiently accurate for orbit determination. However, it is not optimum in all respects. For instance, it would be an improvement to replace the conventional second detector in the receiver with a synchronous detector, or, even better, to remove this detector completely. This could be accomplished by locking the local oscillators in the receiver to the frequency standard. By a suitable choice of local oscillator frequencies, the "difference" frequency at the receiver output could be adjusted to fall within the tuning range of the tracking filter. The net result of this change would be to improve the S/N capability of the receiving system by at least 3 db.

Another modification that may be useful would be to place the present tracking filter, which operates at a relatively low frequency, with an RF servo loop similar to that used in the Microlock system. This would provide some added flexibility by increasing the tuning range of the tracking loop. It may also reduce costs. It is not clear, however, which approach produces the better results. Plans are now being formulated to compare the amount of phase jitter in the voltage controlled oscillators used in each system, and these results will probably be the most important factor in determining which method is to be used.

Other changes that are being considered are adjustment of the "counting time" over which each measurement is made. It will probably be increased, the extent of the change depending on the outcome of error analyses which are planned. Also some revisions of the magnetic tape operation are being studied, because "high" frequency flutter components (>50 cps) have prevented the tracking filters from following the Doppler signal in playback as closely as is desired.

Considerable thought is being given to the development of equipment engineered specifically for this purpose. This will be especially important for navigational instruments where reliability, size, weight, cost, and power requirements will be critical. The first thought that arises in this connection is that the operational equipment must be transistorized and modular.

It is expected that experience will decide the best answers to the above questions, and will certainly dictate the detailed form of equipment for each specific application of a satellite Doppler navigation system.

V. Acknowledgment

The basic idea of making Doppler measurements in the manner described was first developed by a group headed by L. G. DeBey at the Ballistic Research Laboratories, Aberdeen Proving Grounds, Md. It is a pleasure to acknowledge the help the Applied Physcis Laboratory has received from DeBey and his staff.

The physical equipment that actually constitutes a Doppler receiving station is much more complex than the brief description given here would indicate. Needless to say, innumerable problems had to be solved before the individual components could function smoothly as a practical operational unit. It is not possible to list all of the people at this laboratory who contributed to the success of this project. However, the author would like to mention, in particular, the work done by H. H. Elliott who designed many of the special circuits, *e.g.*, the Doppler preset counter and gates, and who was responsible for solving many perplexing practical problems in making the various units operate in unison.

Applications of Doppler Measurements to Problems in Relativity, Space Probe Tracking, and Geodesy*

ROBERT R. NEWTON†

Summary—**This paper begins with a discussion of the precision with which the Doppler shift in the signal received from a space vehicle can be measured, using existing atomic frequency standards on the ground, and a proposed transponder system on the vehicle. Applications of Doppler methods to measuring the gravitational redshift, to tracking space probes and measuring certain astronomical constants, and to geodesy, making use of the precision of atomic standards, are then presented. The conclusion drawn is that Doppler methods can improve upon existing accuracy in these areas.**

Introduction

GUIER and Weiffenbach[1] have described the general way in which Doppler signals from a near Earth satellite can be used either to determine the satellite orbit or to fix the position of the receiving station. Newton[2] has described how the same general method can be extended to the Doppler tracking of an interplanetary vehicle. This paper will describe some possible further extensions of the method.

Let us first point out what seems to be the central feature of the method, namely, that we attempt to use all of the information contained in the Doppler signal from a moving transmitter. This is quite different from another method frequently used in the Doppler tracking of satellites, in which the Doppler is used only to yield the time and range at nearest approach to the receiver (and sometimes the speed).[3]

While many uses of Doppler signals from space vehicles can be developed at the level of accuracy contemplated in the works cited,[1,2] the usefulness of the general method can clearly be enhanced by an increased level of accuracy. Accordingly, we shall first discuss a

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† Applied Physics Laboratory, The Johns Hopkins University, Silver Spring, Maryland.

[1] W. H. Guier and G. C. Weiffenbach, "A satellite Doppler navigation system," this issue, p. 507.

[2] R. R. Newton, "Tracking objects within the solar system using only Doppler measurements," in "1959 Proceedings of the Tenth Congress," International Astronautical Federation, London, in press.

[3] Incidentally, this method has often been misapplied: Let $\boldsymbol{\rho}$ be the vector from receiver to transmitter, with $\rho = |\boldsymbol{\rho}|$; then the classical Doppler frequency is proportional to $\dot{\rho} = \dot{\boldsymbol{\rho}} \cdot \boldsymbol{\rho}/\rho$. The minimum range is inferred from the derivative of this at the instant when $\dot{\rho} = 0$; the derivative of the Doppler at this instant is proportional to $\ddot{\rho} = (\ddot{\boldsymbol{\rho}} \cdot \boldsymbol{\rho} + \dot{\boldsymbol{\rho}} \cdot \dot{\boldsymbol{\rho}})/\rho$. It has been customary to neglect the first term in the numerator, thus making the assumption of unaccelerated motion, even though the first term can be quite appreciable. For example, with a satellite in a circular orbit at 700 km, at a minimum range of 800 km, the first term is about one-seventh of the second term.

method of exploiting the precision of atomic frequency standards. Following this, we shall discuss several applications: to measuring relativistic effects, to tracking interplanetary vehicles and concurrently improving astronomical data, and to making geodetic measurements.

USE OF ATOMIC STANDARDS IN DOPPLER APPLICATIONS

The most obvious way to use atomic frequency standards in space work is to place an atomic clock in the payload, and a program of developing an atomic clock to be flown in a satellite has been announced.[4] We shall describe an alternate method, which can probably be realized sooner and at less expense. This method is inferior in some applications to orbiting an atomic clock, and superior in others.

The alternate method is simple in principle. We start with an Earth-based transmitter, controlled in frequency by an atomic clock, which illuminates the space vehicle. The vehicle receives the signal, the frequency of which is now altered from the transmitted frequency by a Doppler shift, and phase-locks to it an on-board oscillator. The signal from the oscillator is then retransmitted, perhaps with a slight frequency shift for convenience, and the signal is received at an Earth-based station, again modified by Doppler. The resultant of the two Doppler shifts, one occurring with the outgoing signal, and one with the incoming signal, is then extracted by beating the final received frequency against an atomic standard.

There are several possible variants on this basic method. In some applications, we might want to have the Earth-based transmitter and receiver at different locations. This would require having two separate frequency standards, or else a reliable method of sending a standard frequency from one location to the other, and would limit the accuracy to the accuracy of the standards used, or to the accuracy in time and frequency of the ground communication link. In other applications, we would have the Earth-based transmitter and receiver at the same site. This would allow using only a single frequency standard, and it would limit the accuracy only by the frequency drift of the standard during the transit time of the radiation. For satellite work, the transit time would rarely exceed a quarter of a second, and the accuracy could be appreciably greater than the nominal accuracy of the frequency standard.

At this point, we should define more carefully what we mean by the accuracy of a frequency standard. We do not mean the accuracy with which the number of cycles per second of Ephemeris Time is known. If the standard is good enough to yield good Doppler data, then the precision in our knowledge of its absolute frequency is not a limiting factor. Also, we do not mean shifts in frequency which may occur over a few minutes of time. In the applications which we shall consider, observations will be made over a considerable period of time, long enough so that reasonable short-term fluctuations in frequency can be smoothed out statistically. What we do mean is the precision with which a transmitted frequency, averaged over perhaps an hour, can be compared with a received frequency, of almost equal value, and likewise averaged over perhaps an hour, with the precision expressed as a fractional part of one of the two frequencies compared.

With this definition of accuracy, we may expect the accuracy to be about 10^{-10} when different atomic standards are used for the transmitted and received frequencies, and considerably better than 10^{-10} when a single standard is used and when the transit time of the radiation is less than a second.

To analyze the meaning of accuracy in the latter case, let τ be the transit time of the radiation, and let $\epsilon(t)$ be the fractional deviation of the frequency standard at time t from its average over a long time. Then the error in measuring the Doppler frequency at time t is

$$\epsilon(t) - \epsilon(t - \tau) \approx \tau(d\epsilon/dt).$$

The significant quantity is the average (not the rms value) of this over some time interval, from 0 to T, for example. This average is

$$(\tau/T)\left[\epsilon(T) - \epsilon(0)\right].$$

The quantity in brackets will be of the order of 10^{-10} for an atomic standard, and τ/T can readily be of the order of 10^{-3}, so the accuracy for the case in question should be of the order of 10^{-13}. For a good crystal frequency standard, the corresponding accuracy should be better than 10^{-11}.

Another variant on the basic method concerns whether we have an almost constant frequency in the original transmission, or arrange to have an almost constant frequency in the signal received by the vehicle. If we know reasonably well, in advance of a particular experiment, the orbit of the vehicle and the location of the transmitter, we can alter the transmitted frequency by the amount of a predicted Doppler shift. In this way, the frequency received at the vehicle will vary only by the uncertainty in predicting the Doppler, and the stability of the transponder will be greatly enhanced. The price of doing this is a lowered stability in the original transmitted frequency; but it is probable that a net improvement can be effected, particularly when using a single standard for transmission and reception.[5]

Improved precision in frequency is, of course, meaningless unless the performance of all the other components in the system can be made compatible with the frequency performance. In particular, we must be able to measure accurately the time at which the Doppler frequency has a given value. The required precision in time is related to the maximum time derivative of the Doppler frequency. If ν is the transmitted frequency, and ρ is distance from receiver to transmitter, the time derivative of the Doppler frequency is $(\nu/c)\ddot{\rho}$. The time error δt compatible with an error, which we shall write as $\epsilon\nu$, in Doppler frequency is

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[4] For example, see the column "Miscellany" in *Physics Today*, vol. 12, p. 73; September, 1959.

[5] The general idea of using a transponder in a satellite to eliminate the on-board frequency standard must certainly have been suggested independently many times, although we are not aware of any references in the literature. The idea of modulating the transmitted frequency so as to keep the frequency constant at the satellite was suggested to the writer by L. J. Rueger of this Laboratory, and seems to us to have potentially great value.

$$\delta t = \epsilon\nu \div (\nu/c)\ddot{\rho} = \epsilon c/\ddot{\rho}.$$

The minimum δt required occurs when $\ddot{\rho}$ is a maximum. The value will clearly depend upon the application; to get a typical value, if $\ddot{\rho} = 10$ m/second^2, and $\epsilon = 10^{-11}$, $\delta t = 3 \times 10^{-4}$ seconds. This precision is possible with the use of an atomic standard.

The most serious apparent problem in the way of utilizing the available precision is that of correcting the observed Doppler frequency for fluctuations in atmospheric and ionospheric refraction. While many people have reported large sporadic variations in the frequency received from a satellite, amounting sometimes to more than ten cycles at 100 mc, Weiffenbach[6] has observed signals in which such variations are certainly less than a few tenths of a cycle at 100 mc. If we can restrict the data used, for research purposes, to observing periods of such ionospheric quality, we may need to make the refraction corrections to an accuracy of only about one per cent. These corrections can perhaps be made by using more than one frequency, inferring the refraction from the measured dispersion, or by measuring the Faraday rotation and estimating refraction from the rotation.

MEASUREMENT OF RELATIVISTIC EFFECTS

For a near Earth satellite, $\dot{\rho}/c$ is of the order of 10^{-5}. Relativistic effects in the Doppler shift are of the order of $(\dot{\rho}/c)^2$ times the frequency, or $\nu \times 10^{-10}$, where ν is the transmitted frequency. If the accuracy of measuring the Doppler frequency is $\epsilon\nu$, we cannot measure relativistic effects unless ϵ is appreciably less than 10^{-10}.

The most interesting relativistic effect is probably the gravitational red shift. Regardless of the accuracy available, the red shift cannot be measured by a transponder system, since the gravitational effect cancels on the outgoing and return legs. In order to measure the red shift, it is apparently necessary to have a stable transmitter on the satellite whose frequency is known in the satellite frame of reference.

Singer,[7] in his proposal for a measurement of the red shift, and Badessa, Kent, and Nowell,[8] in presenting an ingenious method of cancelling the first-order Doppler shift, apparently assume that an atomic standard, calibrated on the ground, will keep the same frequency, in local units, when satellite-borne. Considering the rigors of launching, and other uncertainties, it seems undesirable to have the accuracy of the experiment rest upon this assumption; it is better to arrange that the standard frequency be measured in flight, using a method which can separate measurement of the frequency from measurement of the red shift.

Accordingly, we suggest carrying a transponder system together with an atomic frequency standard in a satellite. Using the transponder system, we can determine the satellite orbit; the only assumption which this requires about the gravitational red shift is that the shift changes sign according to whether radiation is traveling with or against the field. With the orbit thus determined, study of the Doppler shift in radiation controlled by the atomic standard gives the combined effect of the red shift and a shift in the standard frequency. If the orbit is circular, both of these shifts are constant, and we cannot separate them; however, if the orbit is highly eccentric, we can hope to deduce the dependence of the total shift upon gravitational potential, and hence to separate the two effects.

INTERPLANETARY TRACKING AND ASTRONOMICAL DATA

In work previously cited,[2] the accuracy of tracking the early part of a space probe trajectory by Doppler methods has been estimated at about one part in 10^6 of the orbit parameters, at the end of about half a day of tracking. By going to atomic frequency standards, we can expect to increase the precision to about one part in 10^7 for short-term tracking, and probably better for long-term tracking.

Since the transit time of the radiation is large, the method of using a single frequency standard probably is no more accurate than a method using two standards, although it may be more convenient. It should still be advantageous to premodulate the signals so that the frequency at the space probe is nearly constant.

One result of increased precision should be an increased tracking range, by virtue of decreasing the bandwidth required: the bandwidth can be decreased, in principle, down to the uncertainty in the value of frequency to be received. If scintillations in frequency produced by refraction do not impose a larger limit, we can work at, for example, 0.01 cycle bandwidth at 100 mc, and thus increase the range, on a conservative basis, from 50 million kilometers to perhaps 500 million kilometers, for a few watts of transmitted power from the probe. Whether this limit can be reached in the presence of ionospheric noise remains to be seen.

If the precision in long-term tracking of a space probe can be increased to one part in 10^8 for the orbit elements, one result may be to eliminate the need for a separate midcourse navigation system on board a planetary probe. A Mars probe, for example, could be placed at a point near Mars with a precision of about two miles, accurate enough to start the re-entry phase of the trajectory without the need of an on-board system to track the planet up to this point. Because of the fast reactions required, if for no other reason, a separate system will presumably be needed from this point until the landing.

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[6] G. C. Weiffenbach, private communication.

[7] S. F. Singer, "Application of an artificial satellite to the measurement of the general relativistic 'red-shift'," *Phys. Rev.*, vol. 104, pp. 11–14; October 1, 1956.

[8] R. S. Badessa, R. L. Kent, and J. C. Nowell, "Short-time measurements of time dilation in an earth satellite," *Phys. Rev., Lett.*, vol. 3, pp. 79–80; July 15, 1959.

Another result of increased precision will be an improvement in the values of certain astronomical constants. All quantities which affect the motion of the vehicle, the motion of the observer, or the propagation of radiation between the two, affect the Doppler frequency. Thus, by measuring the Doppler, we can get a measure of these quantities, with an accuracy which depends upon the sensitivity of the Doppler to them, provided that we make enough measurements to separate out the different ways in which the different quantities affect the Doppler. Among such quantities will certainly be the mass ratios of Sun, Earth, and Moon, the values of their mean distances in terms of laboratory units of distance, and the velocity of light in interplanetary space. If the tracking extends over a few months in time, which is a significant part of the orbital periods of Mercury and Venus, we may even obtain improved values of their masses by analyzing perturbations upon the space probe orbit.

It is perhaps surprising that we can determine geodetic data of currently useful quality from tracking an interplanetary vehicle. However, the component of velocity which the receiving station has because it is carried upon the surface of the rotating Earth (and, consequently, its geocentric coordinates), enters directly into determining the Doppler shift. If the shift is measured to a precision of $\epsilon\nu$, the velocity is measured to a precision of ϵc, or about 0.003 m/second for $\epsilon = 10^{-11}$. Rotational velocity at the equator is about 460 m/second; thus, for a station at the equator, we can determine the local radius of the Earth and the east-west position of the receiver within about 40 m. This exceeds the present accuracy for some parts of the Earth.

It is important to realize, first, that these geodetic data are obtained from a system which is completely independent of any anomalies in the Earth's gravitational field (since it does not significantly depend upon Earth's gravitation at all), and, second, that the data obtained are directly the coordinates of the receiving station in the coordinate system whose origin is the terrestrial center of mass, and one of whose axes is the instantaneous axis of rotation of the Earth. Measurements made in this way are therefore independent of the traditional methods of geodesy, being based upon independent principles, and therefore can form a valuable supplement to geodetic data.

Geodetic measurements having this same advantage can, however, be obtained with greater accuracy using Doppler data from a high altitude Earth satellite. This brings us to a model for a geodetic satellite.

A GEODETIC SATELLITE

The figure of the Earth enters into determining the Doppler shift in the signal from a satellite in two independent ways. First, the figure, in conjunction with the density distribution, determines the gravitational field in which the satellite moves. Second, the figure, in a fashion which is purely geometrical and independent of the density distribution, determines the velocity of the receiving station, which is part of the relative velocity between transmitter and receiver. It is desirable to make geodetic measurements which separate these effects, perhaps by eliminating one of them.

Use of a space probe for geodetic work accomplishes this purpose by eliminating the effects of the Earth's gravitation, but at the expense of decreasing the sensitivity of Doppler measurements to the receiver coordinates. A near Earth satellite gives greater sensitivity to the coordinates, but at the expense of making the orbit sensitive both to local gravitational anomalies, including the time-dependent effects produced by tides, and to variations in drag.[9] A compromise is desirable: for geodesy we should use the lowest-altitude orbit which is unaffected, to the desired accuracy, by gravitational anomalies or by fluctuating drag, within the time period needed for the observations. This time will be of the order of a few orbital periods.

We do not know what the optimum altitude for this geodetic satellite should be. We shall assume that the altitude is one Earth radius; this altitude is unlikely to be too low, so that we shall be conservative in using this altitude when we estimate the accuracy of determining coordinates.

It is difficult to estimate the accuracy for a general orbit and a general observing position. We have made an estimate for the particular case of a circular equatorial orbit at an altitude of one Earth radius, and an observer near the equator. The method of calculation is similar to the standard method outlined, for example, by Newton,[2] except that we assume here that the mean value of the population from which a noise sample is drawn is not zero, but increases uniformly from zero to $2\epsilon\nu$ during a single pass. This puts both a drift and a bias into the observations of frequency. σ_R, the error in the local radius of the Earth, and σ_L, the error in longitude, become:

$$\sigma_R = 2 \times 10^{11}\epsilon \text{ meters},$$

$$\sigma_L = 5 \times 10^{11}\epsilon \text{ meters}.$$

The error in measuring latitude is approximately $\sigma_R \sec \theta_{\max}$, where $\theta_{\max}$ is the maximum elevation of the satellite during the pass. Unless $\theta_{\max}$ is too close to 90°, the error in latitude is comparable to σ_R.

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[9] D. G. King-Hele, "Determination of air density and earth's gravitational field from the orbits of artificial satellites," in "1959 Proceedings of the Tenth Congress," International Astronautical Federation, London, in press.

For $\epsilon = 10^{-11}$, then, we can find the geocentric coordinates of the observer to an accuracy of a few meters from a single pass, provided that accuracy is limited only by the accuracy with which a Doppler frequency can be measured. Since we can probably make $\epsilon < 10^{-11}$, and can use more than one pass, we can hope to measure coordinates to a meter or less, again with the assumption that the problems other than frequency measurement can be solved. We have already discussed some of these other problems.

In using this satellite, we shall probably find it helpful to use both the method of a single standard, with receiver and transmitter at the same location, and the method of separate locations. The former gives greater precision, but the latter yields the relative coordinates of transmitter and receiver directly, and should facilitate some studies. It should also be possible to make measurements from a surface vessel, although with some loss of accuracy, thus enormously increasing the geodetic coverage of the globe.

We should almost certainly use the method of modulating the originally transmitted frequency, in order to maintain constant frequency at the satellite.

It is, of course, desirable to make other accurate geodetic measurements using methods which involve gravitational anomalies, so that these anomalies can be studied.

Acknowledgment

The writer is grateful for helpful suggestions made by many of his colleagues at the Applied Physics Laboratory, particularly G. B. Bush and L. J. Rueger.

A Doppler-Cancellation Technique for Determining the Altitude Dependence of Gravitational Red Shift in an Earth Satellite*

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Summary—A cancellation technique permits measurement of the frequency of a source moving relative to an observer without the obscuring effect of first-order Doppler shifts. The application of this method to a gravitational red shift experiment involving the use of an earth satellite containing a highly stable oscillator is described. The rapidity with which a measurement can be made permits the taking of data at various altitudes in a given elliptical orbit. Tropospheric and ionospheric effects upon the accuracy of results are estimated.

Introduction

According to the General Theory of Relativity, the time kept by a "proper" clock (or the frequency of a standard oscillator) is affected by the motion of the clock and by its local gravitational potential. The latter effect, known as the "gravitational red shift," is expected to result in variations of only 7 parts in 10^{10} in the rates of clocks in the vicinity of the earth. Attempts to obtain experimental verification have been stimulated by the development of highly stable frequency standards and the availability of earth satellites in orbits that provide a significant variation in gravitational potential.

In 1956, Singer[1] proposed the measurement of gravitational red shift by an experiment involving comparison of the time kept by a stable clock in a satellite with that of an identical clock on earth. Since then, several versions of his experiment have been suggested. Basically, all of them involve the use of an atomic clock, or equivalent standard, from which, by frequency division, a series of precise timing pulses can be derived. These pulses would be transmitted as a modulated carrier over a one-way link from a satellite to earth, where they would be compared with a similar, locally generated timing waveform to determine the frequency difference. The frequency of the received pulses is directly influenced by the rate-of-change of path lengths (first-order or radial Doppler shift). This effect is typically four or five orders of magnitude larger than the gravita-

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[1] S. F. Singer, "Application of an artificial satellite to the measurement of the general relativistic 'red shift'," *Phys. Rev.*, vol. 104, pp. 11–14; October, 1956.

tional red shift. Consequently, data must be acquired over a sufficiently long time interval (probably several days) to insure that path-length measurement errors are negligible. At the completion of this experiment, a single value is obtained for the average frequency shift over many orbital periods. Thus it is not possible by this technique to determine the effect of altitude variations.

In order to obtain the altitude dependence of gravitational red shift, the measurement interval must be shortened to a small fraction of the orbital period. However, path-length variations make the performance of a short-time measurement in a one-way propagation experiment virtually impossible. But two-way propagation experiments offer definite possibilities. Engineers are quite familiar with the doubling of the first-order Doppler shift, which occurs in transmission of signals to and from a radar target. If it were possible to reverse the sense (*i.e.*, direction) of one of these frequency shifts by a mixing operation, then essentially complete cancellation of first-order Doppler shift could be effected. In the short-time measurement system[2] that will be described, such a technique permits data for a single altitude to be obtained in an interval of 60 seconds or less. The following significant advantages result.

1) By employing an elliptical orbit, frequency differences at various altitudes can be observed, and hence the variation of frequency with gravitational potential can be obtained.
2) If the measurement is repeated when the satellite is at a particular altitude (*e.g.*, at perigee), a slow frequency drift between satellite and earth clocks can be determined and removed. Since the quantity of interest is the *variation* in frequency difference as a function of altitude, a fixed offset frequency between clocks is inconsequential. Therefore, quartz-crystal oscillators can probably be employed.
3) Continuous integration or counting of pulses over many orbital periods is not required. Skipped cycles or temporary failures would not invalidate the experimental results.

The Short-Time Measurement Principle

The principle employed in a short-time measurement will be illustrated by a simplified analysis. In the interest of simplicity, the measurements are assumed to take place at an earth pole (O in Fig. 1). (An outline of the derivation of the following expressions is given in the Appendix.)

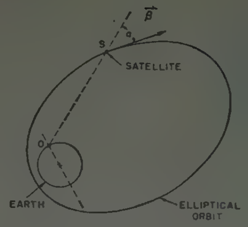

Fig. 1—Geometry of the measurement problem.

1) A transmitter (located at point O) of frequency f transmits to the satellite.
2) The frequency of the signal received at the satellite is

$$f' = f\frac{1 - \beta \cos \alpha}{(1 - \beta^2 + 2\phi)^{1/2}}, \tag{1}$$

where β is the magnitude of the satellite velocity (measured in the coordinate system of the observer at O) at the instant of reception, normalized in terms of the velocity of light; α is the angle (measured in the same coordinate system) between the propagation path O–S and the velocity vector $\vec{\beta}$; ϕ is the (positive) difference in gravitational potential between points S and O, normalized in terms of the square of the velocity of light.
3) The satellite contains an oscillator of frequency $2f$ (measured on the satellite). The received signal frequency f' is subtracted from $2f$ by mixing within the satellite to yield a frequency difference

$$f'' = 2f - f'. \tag{2}$$

4) The mixture signal f'' is transmitted back to earth. (Reception and retransmission at the satellite are assumed to be virtually instantaneous.)

The frequency of the received signal is measured as

$$f''' = f''\frac{(1 - \beta^2 + 2\phi)^{1/2}}{1 + \beta \cos \alpha}. \tag{3}$$

Substitution for f'' in (3) yields

$$f''' = f\left[2 - \frac{1 - \beta \cos \alpha}{(1 - \beta^2 + 2\phi)^{1/2}}\right]\left[\frac{(1 - \beta^2 + 2\phi)^{1/2}}{1 + \beta \cos \alpha}\right]. \tag{4}$$

By means of a power-series expansion, retaining second-order terms in β and first-order terms in ϕ, we obtain

$$f''' \approx f(1 + 2\phi - \beta^2).$$

The first-order Doppler term has been cancelled, and the received signal frequency differs from the ground transmitter frequency by

$$\Delta f \approx f(2\phi - \beta^2). \tag{5}$$

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[2] R. S. Badessa, R. L. Kent, and J. C. Nowell, "Short-time measurement of time dilation in an earth satellite," *Phys. Rev. Lett.*, vol. 3, pp. 79–80; July, 1959.

The two terms in (5) are apparently inseparable experimentally. The first term, $2\phi f$, is the desired gravitational frequency shift. The second term, $-\beta^2 f$, is the well-known second-order, or transverse, Doppler shift of special relativity. This term, which has been verified experimentally, can be calculated easily and subtracted from the measured frequency difference to obtain the residual gravitational effect.

The frequency difference in the general case of an observer not located at the pole (Fig. 2) is given by

$$\Delta f = f[2\phi - (\beta_s^2 - \beta_0^2) - 2\beta_0\beta_s \cos(\alpha_2 + \gamma_2) - \beta_0^2(1 + r \sin \gamma_2)], \quad (6a)$$

where β_s, β_0, α_2, and γ_2 are as shown in Fig. 2, measured in a nonrotating coordinate system with origin at the earth's center and r is the range, normalized in terms of the earth's radius. The term in (6a) containing the product $\beta_0\beta_s$ is of the order of 10^{-10}, and cannot be neglected in the present experiment. This term must be retained, and appropriate values substituted. Present orbital computational accuracy is more than adequate for this purpose.

SATELLITE EQUIPMENT

As we have indicated, the operations to be performed in the satellite are as follows:

1) Amplify a received signal of frequency f'.
2) Mix the received signal with that of the satellite "clock" to obtain $2f-f'$.
3) Amplify and transmit a signal of frequency $2f-f'$.

In order for the cancellation of first-order Doppler shifts to be effective, the signals received and transmitted at the satellite must be very nearly of the same frequency. One way of avoiding the ensuing "spillover" problem is to employ pulsed waveforms. The received signal pulses, 1–10 μsec long, are stored briefly in a delay line or equivalent device. After the reception of each pulse, the receiver is turned off and the transmitter turned on. Fig. 3 is a block diagram of the satellite equipment required to instrument this technique. This configuration is intended only for purposes of illustration. An operating frequency range of 500–1000 mc, which is expected to fall between regions of severe ionospheric and tropospheric propagation anomalies, has been tentatively selected. A factor in the final choice of an operating frequency is the desirability of employing transistor circuitry. For ranges up to 10,000 miles, an RF output of approximately 20 mw (average), 0.4 watt (peak) is required from the satellite transmitter.

Although duplication of units is not indicated in the block diagram, it is particularly desirable to include more than one crystal frequency standard in the satellite, to permit independent measurements and to improve system reliability. The use of three crystal oscillators in a common temperature-stabilized oven is suggested. The crystals, which could be offset from one another in frequency for identification purposes, might be selected with different known temperature coefficients. Then variations in satellite temperature, which could possess a deceiving similarity to the orbital period of the satellite itself, will affect the three oscillators in varying degrees. Thus, a means for separating relativistic frequency shifts from temperature-dependent effects is provided. Preliminary power and weight estimates for the satellite equipment, including the weight estimate for batteries, are 3 watts and 35 pounds.

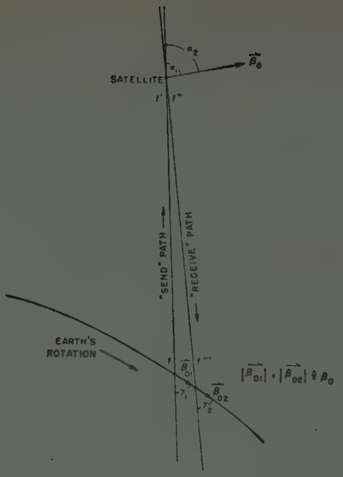

Fig. 2—Geometry of the two-way propagation experiment.

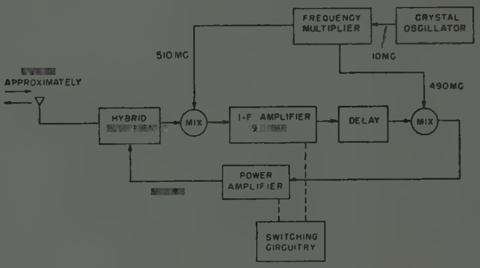

Fig. 3—Illustrative satellite-equipment configuration.

The stability of the crystal oscillators is of paramount importance and will be discussed separately. The overall phase shift through the receiver and transmitter should remain stable within 30° over the measurement interval 20–60 seconds, in order not to degrade the measurement accuracy which should be determined essentially by the short-term stability of the crystal oscillators.

GROUND EQUIPMENT

An illustrative block diagram of the major portion of the required ground equipment is shown in Fig. 4. It functions as follows:

1) A signal of frequency f is transmitted.
2) A signal of frequency f''' is received from the satellite and compared in frequency with f by means of a mixer and a synchronous detector.
3) The frequency-difference signal is recorded on tape together with a fixed-frequency reference signal.

The transmitted and received signals both consist of series of pulses (each 1–10 μsec long). Time-delay discrimination is provided in the receiver for the rejection of transmitter feed-through signals and other undesired signals. In addition, a frequency offset of 1 part in 10^6 is purposely introduced between the ground and satellite frequency standards. This insures a small frequency separation between the signals transmitted and received on the ground, thereby facilitating resolution of the satellite signal from clutter by means of narrow-band filtering, without materially affecting first-order Doppler cancellation.

The coherent waveform and narrow-band filtering (±1 cps) employed in the ground receiver make possible the low requirement on satellite transmitter power (20 mw, average). Doppler tracking is unnecessary because the frequency of the signal received on the ground is virtually independent of first-order Doppler shift. However, the first-order Doppler shift is present in the signal received on the satellite. For simplicity, broad-band amplification, rather than Doppler tracking, is employed there, at the expense of increased ground-transmitter power. For a 10 meter^2 antenna on the ground, the power required to yield a clean signal on the satellite is approximately 500 watts (average), 10 kw (peak).

The ground station requires means for storage, as well as immediate processing, of the receiver output signal. The desired information is contained in the small variations in output frequency with orbital position. Since orbital position will be known as a function of time, a recording of the output signal together with a time signal will suffice.

Processing of the data involves measurement of a small frequency difference between two signals. This is best accomplished by measuring the total phase difference accumulated in a specific time interval. For example, a frequency difference of 0.1 cps in a 60-second interval results in a total accumulated phase difference of 2160°. This can be measured precisely by manual phase-tracking techniques, with the use of an oscilloscope as a visual tracking indicator.

OSCILLATOR STABILITY

The frequency standards used in this experiment must meet two separate stability requirements. First, all random and periodic frequency excursions must be at least an order of magnitude smaller than the total effect that is to be measured. Second, any long-term frequency drift in one orbital period should not exceed 20 to 30 per cent of the total desired effect. Perhaps the reasons for these stability requirements can best be illustrated by plotting a set of hypothetical data that might be obtained in this experiment. Fig. 5 shows a plot of the hypothetical frequency difference against time. For simplicity, it is assumed that data are acquired alternately at apogee and at perigee for several successive sightings. The clusters of points at times A, B, C, and so on, represent several readings of frequency difference obtained during the intervals when the satellite is visible. The frequency difference between the two dashed lines for apogee and perigee readings is expected to be approximately 4 parts in 10^{10}. Therefore, the undesired random and periodic frequency variations of the oscillators that are indicated by the dispersion of the clusters must be no greater than a few parts in 10^{11}.

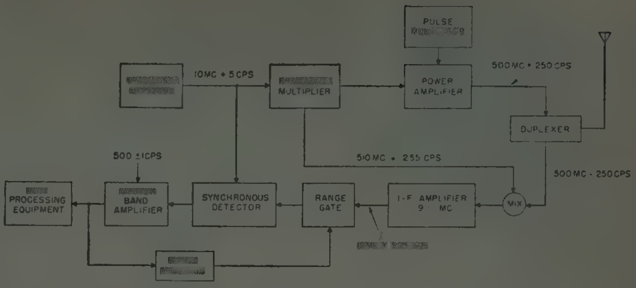

Fig. 4—Illustrative ground-equipment configuration.

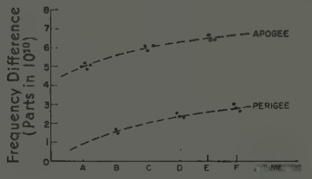

Fig. 5—Hypothetical data to illustrate oscillator stability requirements.

The slope of either of the dashed lines in Fig. 5 represents the difference between the long-term drifts of the oscillators, and it is clear that a monotonic drift per orbital period of from 20 to 30 per cent of the total effect can easily be tolerated. For the proposed experiment, a drift of 1 part in 10^{10} in a few hours, or approximately 1 part in 10^9 per day, is permissible. A small fixed offset frequency between the oscillators would merely shift the base line in Fig. 5, and therefore would have no effect upon the desired measurement.

Actual measurements taken on commercially available quartz-crystal oscillators indicate that these oscillators essentially meet the stability requirements. Table I shows some data obtained in a comparison test. The 1-mc output signals of two crystal oscillators were multiplied up and mixed down in such a way as to magnify their frequency difference by a factor of 1000. The resulting beat frequency was then determined by means of a counter operating over periods of 10 and 100 seconds. The last digit of the counter output represents parts in 10^{10} for the 10-second count, and parts in 10^{11} for the 100-second count. In either case, the counter has a starting error of one count. Thus, the differences indicated between the two oscillators may be entirely accounted for by the counter error. The oscillator error is shown to be no greater than 1 part in 10^{11} when it is averaged over successive 100-second intervals.

TABLE I

FREQUENCY COMPARISON DATA FOR TWO HERMES MODEL 101C 1-MC CRYSTAL FREQUENCY STANDARDS

10-Second Counting Interval; Total Elapsed Time, Approximately 45 Minutes.	100-Second Counting Interval; Total Elapsed Time, Approximately 15 Minutes
93101	31005
93100	31006
93100	31006
93100	31005
93100	31006
·	31006
·	31006
·	
·	
93100	
93100	
93101	
93100	
93101	

The oscillators that were tested employed vacuum-tube circuits; data are not available for a transistor version that is now being developed. However, the stability is believed to be determined primarily by the crystal and its oven, and not by the oscillator circuitry. Therefore no degradation in stability is expected to result from a change to transistors.

In order to transfer the oscillator stability to a transmitter operating in the UHF range, a chain of frequency multipliers is required. Two methods of frequency multiplication, both of which use variable-capacitance diodes as the multiplying elements, are being considered. Leeson and Weinreb,[3] in a detailed study of low-order frequency multipliers with nonlinear reactances, demonstrated that frequency doublers with only 1-db loss could be constructed. A chain of doublers that are now being developed to go from the oscillator frequency to the transmitter frequency is expected to yield a loss of only 10–15 db. The second approach to the frequency-multiplication problems is the use of one or two large frequency jumps. However, experimental results thus far indicate that in multiplying by a factor of 100, the loss is approximately 40 db.

PROPAGATION EFFECTS

As we have stated, the success of the described measurement depends upon the cancellation of the first-order Doppler shift, an effect that is four or five orders of magnitude greater than the gravitational red shift. For adequate cancellation, it is essential that random fluctuations in the effective path lengths upward and downward be small, or that they be well correlated. Phase fluctuations along the propagation paths are known to be caused by turbulent irregularities in the troposphere and in the F_2 region of the ionosphere. Extensive discussions of these and related effects have been published.[4–10]

It is desirable to select an operating frequency in a region in which the total rms phase error that is expected to arise from propagation anomalies falls well within the total permissible phase error of the experiment. If the frequency shift is computed by observing the total accumulated phase difference in a specified time interval T, then the resulting frequency error cannot exceed $\Delta\psi/\pi T$ cps, where $\Delta\psi$ is the total uncancelled phase fluctuation. At 500 mc, an rms phase error of 0.5 radian results in a 1 per cent error in frequency for a 60-second measurement interval. Such an error would be acceptable, since it is less than the value of the expected oscillator instability.

Tropospheric fine structure is measured by variations in the dielectric constant. The rms one-way phase fluctuation obtained by Muchmore and Wheelon[4] is given by

$$\psi_{\rm rms} \approx \frac{(2l_0L)^{1/2}\pi\Delta e}{\lambda}, \qquad \text{for } L \gg l_0, \tag{6b}$$

where l_0 represents the scale of the irregularities (probably 200 feet); L is the propagation path length through the turbulent medium (taken to be 20,000 feet); Δe is the rms deviation of the dielectric constant from unity (probably 10^{-6}); and λ is the wavelength. At a frequency of 1000 mc, the one-way phase fluctuation is computed as approximately 0.6° rms. This value is believed to be representative for a turbulent mass of (optically) clear air. In the presence of rain clouds, the interiors of which are very turbulent and contain large fluctuations in dielectric constant, the situation is somewhat different. If we take as representative values for a rain cloud $l_0 = 20$ feet, $L = 5000$ feet, $\Delta e = 20 \times 10^{-6}$, the one-way fluctuation in phase at 1000 mc is found to be approximately 1.6° rms. Thus, a net phase fluctuation of 2° or 3° rms might be typical for 1000-mc one-way propagation through the troposphere on a cloudy day.

The tropospheric phase-fluctuation measurements of Herbstreit and Thompson,[10] at 1046 mc, appear to be in general agreement with theory. The standard deviation of the phase shift over a 10-mile tropospheric path exceeded 3° fifty per cent of the time, and 5.5° ten per cent of the time. Moreover, the fluctuation period was found to be approximately 1 hour for large fluctuations, with minute-to-minute variations in phase shift of only 0.3° rms. Since the tropospheric phase fluctuation, from (6b), varies directly with the operating frequency, we conclude that the contribution of the troposphere to measurement errors in this experiment is negligible.

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[3] D. B. Leeson and S. Weinreb, "Frequency multiplication with nonlinear capacitors—a circuit analysis," PROC. IRE, vol. 47, p 2076–2084; December, 1959.

[4] R. B. Muchmore and A. D. Wheelon, "Line-of-sight propagation phenomena. I. Ray treatment," PROC. IRE, vol. 43, pp. 1437–1449; October, 1955.

[5] J. L. Pawsey and R. N. Bracewell, "Radio Astronomy," Clarendon Press, Oxford, Eng., pp. 347–351; 1955.

[6] H. G. Booker, "The use of radio stars to study irregular refraction of radio waves in the ionosphere," PROC. IRE, vol. 46, pp. 298–314; January, 1958.

[7] D. F. Martyn, "The normal F region of the ionosphere," PROC. IRE, vol. 47, pp. 147–155; February, 1959.

[8] C. O. Hines, "Motion in the ionosphere," PROC. IRE, vol. 47, pp. 176–186; February, 1959.

[9] "Propagation Factors Affecting Long-Range UHF Radars at High Latitudes," Defence Res. Telecommun. Establ., Ottawa, Ont., Can., Rept. No. 41-1-3; December, 1958.

[10] J. W. Herbstreit and M. E. Thompson, "Measurements of the phase of radio waves received over transmission paths with electrical lengths varying as a result of atmospheric turbulence," PROC. IRE, vol. 43, pp. 1391–1401; October, 1955.

Ionospheric fluctuations are measured by variations in the free-electron density. Turbulent irregularities in the F_2 region are maximum near midnight (local time) and occur in a region roughly 55 km thick at a nighttime altitude of approximately 300 km. The one-way rms phase fluctuation is shown by Booker[6] to be

$$\psi_{\text{rms}} = \left(\frac{h_0 H}{2}\right)^{1/2} r_e \lambda \Delta N, \tag{7}$$

where h_0, H are analogous to l_0, L for the troposphere (taken here to be 1 km and 55 km, respectively); r_e is the classical radius of the electron, 2.8×10^{-15} meter; and ΔN is the rms fluctuation in free-electron density. A value of $\Delta N = 10^{11}/\text{meter}^3$ is believed to be representative of the severe fluctuations encountered during periods of intense sunspot activity. Use of this value yields a phase fluctuation of 0.44 radian rms at 1000 mc. The ionospheric effect, from (7), varies inversely with the operating frequency. The desired relativistic effects, from (5), vary directly with frequency. Therefore, the fractional measurement error caused by ionospheric turbulence will vary inversely as the square of the operating frequency. Recent UHF measurements indicate that the phase fluctuation in propagation through the ionosphere is much less than one radian rms during a daytime period of normal ionospheric activity.

The Doppler-cancelling property of the two-way propagation experiment will effect cancellation of phase (or frequency) fluctuations resulting from propagation anomalies, as well as the radial Doppler shift resulting from the motion of the satellite, to the extent that correlation exists between the fluctuations introduced in the upward and downward traversals of the turbulent medium. The correlation coefficient is still to be estimated, in order to determine the residue of the calculated one-way phase fluctuation that will remain uncancelled in the experiment. It is known that the ionosphere rotates with the earth, and hence the upward- and downward-ray paths pass through essentially the same region of the ionosphere. The round-trip time delay, τ, associated with the largest anticipated range (10,000 miles) is 0.1 second. The value of the correlation coefficient for phase fluctuations displaced up to 0.1 second in time is therefore required. Experimental data[6,9] indicate maximum fluctuation rates of approximately 4 cycles per minute, occurring during periods of magnetic activity. Assuming an exponential correlation function, we have

$$\rho(\tau) = e^{-\omega_0 \tau}, \tag{8}$$

where $\omega_0 = 2\pi/15 = 0.42$ radian/second.

The residual (uncancelled) phase error is then given by

$$\Delta\psi_{\text{rms}} = \psi_{\text{rms}}[2(1-\rho)]^{1/2} \tag{9}$$

which, for small values of $\omega_0 \tau$, is approximately

$$\Delta\psi_{\text{rms}} \approx \psi_{\text{rms}}(2\omega_0\tau)^{1/2}, \qquad (\omega_0\tau \ll 1). \tag{10}$$

For $\tau = 0.1$ second, the quantity $(2\omega_0\tau)^{1/2}$ in (10) is equal to 1/3.5. The Doppler-cancellation process is therefore effective in reducing ionospheric phase fluctuations by a factor of 3.5, or better. The uncancelled phase shift is found to be 7.4° rms at 1000 mc, or 15° rms at 500 mc, for a path length (one-way) of 10,000 miles. Even at the lower operating frequency, the expected error for a 60-second measurement interval is only 0.0014 cps, or 0.4 per cent of the expected maximum red shift.

We conclude that for operating frequencies in the range 500–1000 mc, propagation anomalies will produce a negligible deterioration in the accuracy of the measurement.

APPENDIX

We shall outline a method for computation of the Doppler frequency shift associated with the propagation of electromagnetic signals between points in nonuniform motion in a gravitational field. An approximate result is derived here; an exact derivation has been made by Friedman.[11]

If a gravitational field in space is described in a system of coordinates x_1, x_2, x_3, x_4, where $x_4 = ct$, and t represents "coordinate time," then the relation

$$ds^2 = \sum_{\mu=1}^{4} \sum_{\nu=1}^{4} g_{\mu\nu} dx_\mu dx_\nu \tag{11}$$

holds between each locally measured interval ds and the corresponding coordinate differences, dx_μ. For events occurring at a single point, the locally measured interval is given by

$$ds = c d\tau, \tag{12}$$

where $d\tau$ denotes "proper time" at that point. The quantities $g_{\mu\nu}$ define the metric properties of space (*i.e.*, the gravitational field). For a spherically symmetric distribution of mass about the origin (Fig. 6), the $g_{\mu\nu}$ are given by the Schwarzschild solution to the field equations of general relativity:

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[11] M. H. Friedman, Dept. of Physics, M.I.T., Cambridge, Mass., private communication; October, 1959.

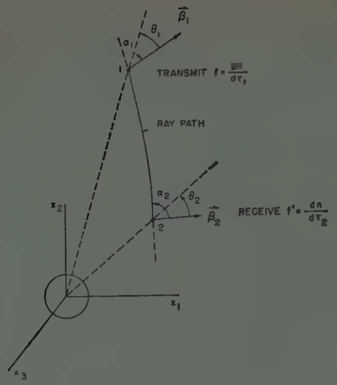

Fig. 6—Geometry of propagation between two points moving in a spherically symmetric gravitational field.

$$\left.\begin{aligned} g_{44} &= 1 + 2\phi \\ g_{4\nu} &= g_{\mu 4} = 0 \\ g_{\mu\nu} &= -\delta_{\mu\nu} + \frac{2\phi}{1+2\phi}\frac{x_\mu x_\nu}{r^2} \end{aligned}\right\} \quad \begin{aligned} \mu &= 1, 2, 3 \\ \nu &= 1, 2, 3, \end{aligned} \tag{13}$$

where $\delta_{\mu\nu} = 1$ $(\mu = \nu)$; $\delta_{\mu\nu} = 0 (\mu \neq \nu)$; r is the distance to the origin from the point x_1, x_2, x_3; and ϕ is the normalized gravitational potential at that point.

Consider a transmitter (Fig. 6) at point 1 in gravitational potential ϕ, whose frequency f is precisely controlled by a resonance of atoms, traveling with respect to the origin with (normalized) velocity β_1. The frequency transmitted may be regarded as some number of events per unit time. As viewed in the x_μ coordinate system, this frequency is given by

$$f_1 = \frac{dn}{dt_1} = \frac{dn}{d\tau_1}\frac{d\tau_1}{dt_1} = f\frac{d\tau_1}{dt_1}\,. \tag{14}$$

Substitution for $g_{\mu\nu}$ and ds in (11) yields

$$\frac{d\tau_1}{dt_1} = [1 + 2\phi_1 - \beta_1^2(1 - \epsilon_1 \cos^2\theta_1)]^{1/2}, \tag{15}$$

in which $\epsilon_1 = (2\phi_1)/(1+2\phi_1)$; θ_1 is the angle between $\vec{\beta}_1$ and the radius to point 1. The observer at point 2 is assumed to utilize a frequency standard of like constitution to that at point 1. By the same method, the frequency received at point 2 (viewed in the x_μ coordinate system) can be related to the frequency measured. Thus

$$f_2 = f'\frac{d\tau_2}{dt_2} \tag{16}$$

and

$$\frac{d\tau_2}{dt_2} = [1 + 2\phi_2 - \beta_2^2(1 - \epsilon_2 \cos^2\theta_2)]^{1/2}, \tag{17}$$

where dt_2 is the (coordinate) time of arrival of the ray at point 2. Since the number of events is preserved,

$$dn = f'\frac{d\tau_2}{dt_2}\,dt_2 = f\frac{d\tau_1}{dt_1}\,dt_1. \tag{18}$$

Therefore

$$f' = f\frac{dt_1}{dt_2}\left[\frac{1 + 2\phi_1 - \beta_1^2(1 - \epsilon_1 \cos^2\theta_1)}{1 + 2\phi_2 - \beta_2^2(1 - \epsilon_2 \cos^2\theta_2)}\right]^{1/2}. \tag{19}$$

The relationship between the coordinate times of reception and transmission is intractable because the ray velocity varies slightly along the path, and the path itself is curved. However, it can be shown that neglect of these effects leads to an error of the order of the satellite velocity multiplied by the gravitational potential, which is only approximately 10^{-14} for earth satellites. For present purposes, the derivation can be completed by a classical computation of the rate-of-change of path length:

$$\frac{dt_1}{dt_2} \approx \frac{1 - \beta_2 \cos\alpha_2}{1 + \beta_1 \cos\alpha_1}\,. \tag{20}$$

Moreover, ϵ_1 and ϵ_2 in (19) can be neglected. Thus,

$$f' \approx f\left[\frac{1 + 2\phi_1 - \beta_1^2}{1 + 2\phi_2 - \beta_2^2}\right]^{1/2}\left(\frac{1 - \beta_2\cos\alpha_2}{1 + \beta_1\cos\alpha_1}\right). \tag{21}$$

ACKNOWLEDGMENT

The authors wish to acknowledge the assistance given by Prof. M. H. Friedman, Prof. J. W. Graham, and V. J. Bates. They would also like to express their appreciation for the invaluable guidance of Prof. J. R. Zacharias.

Attitude Reference Devices for Space Vehicles*

P. E. KENDALL†, SENIOR MEMBER, IRE, AND R. E. STALCUP†

Summary—**As the avenues of space travel broaden, the problem of determining the attitude of the space vehicle becomes more difficult; hence, the need for a compact, reliable sensing device is becoming increasingly important.**

This paper presents the design features of an attitude sensing device which is quite compact and should prove to be reliable. The original objective in the design of the device was to provide a simple means of establishing a vertical reference line from an orbiting space vehicle to a planet under observation. It has been accomplished through the use of two narrow infrared scanning beams. The center line of each beam rotates at constant angular velocity in a plane, the plane of one beam being perpendicular to that of the other. The active portion of the scan is 270° for each beam. When the scanner intersects the planet, a step in the signal level occurs. Position of the beam when the half width of the signal step occurs establishes a plane which contains the center of the planet. Intersection of the two such planes generated by the dual scanner is the direction to the planet local vertical. The assembly is described in detail.

Further, by using analog computer techniques, it is shown that pitch and roll angles about this vertical reference line can be generated. A single degree of freedom gyro can be added to the equipment to produce yaw information in the case of an orbiting vehicle. Finally, a method for measuring altitude is discussed.

INTRODUCTION

As the avenues of space travel broaden, the problems associated with the control of the space vehicle become more complex, and at the same time of considerably more importance. The initial space vehicles have been fairly simple, with a limited amount of instrumentation aboard the vehicle and little importance attached to the orientation of the vehicle. In the succeeding months of this year, and in the years following, the attitude of the vehicle is going to be considerably more important, not only in knowing what it is, but also in being able to control it. For example, if we were photographing the earth from such a vehicle, it would be desirable to have the photographic device looking at the earth most of the time as the vehicle orbits.

The problem of attitude is complicated by a consideration of spinning vehicles. Quite often satellites are launched with a spinning motion applied to provide stabilization and insure a uniform thrust direction. In general, the spin velocity has not been too carefully controlled, but if a man is going to be aboard the vehicle, spin may be undesirable, or spinning at a predetermined, controlled rate may be highly desirable. In considering the problems of devising an attitude reference device which will be operative for spinning and nonspinning vehicles, a device has been chosen which will operate up to spin rates of 60 rpm. In addition, a range of altitudes of 150 to 600 nautical miles has been considered. Actually, higher altitudes could be accommodated by decreasing the beamwidth of the scanner. The vertical reference device described in this paper has been designed so as to require a minimum of space, and to be operative for extended periods of time, in fact for periods of time up to two years or longer. In keeping with the thought that the device should be useful for a broad range of vehicles, the design encompasses the measurement of several vehicle parameters, namely, roll, pitch and yaw about the local vertical, as well as altitude. The latter can be eliminated with a consequent improvement in size and weight. Yaw can also be eliminated with a still further improvement in size and weight. Actually, the requirements of the mission will dictate the amount of equipment to be used. The design is on a building-block basis to permit such utilization. Details of the equipment are described in the following pages.

HORIZON SENSING

Scanning is done by two narrow search beams operating simultaneously. The center line of each beam rotates at constant angular velocity in a plane, the plane of one being perpendicular to that of the other. These planes, hereafter referred to as scanning planes, are fixed, with respect to the vehicle axes, in a position such that their line of intersection coincides with the vehicle yaw axis (vertical reference line).

The manner in which the location of the local vertical is established is described by reference to Fig. 1, which

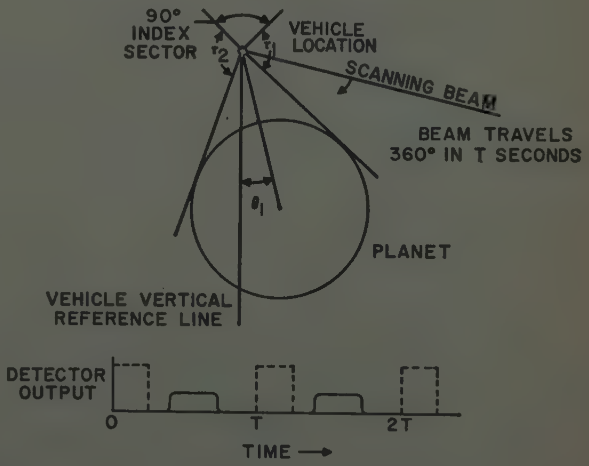

Fig. 1—Operation in one scanning plane.

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shows the geometry of a typical situation in one scanning plane. The cross section of the planet in the scanning plane, lines tangent to the cross section, the vehicle vertical reference line, and orientation of the index sector are shown to explain the appearance of the periodic waveform which is sketched in Fig. 1(b). Dotted pulses are shown for periods when the scanning beam is on the index sector. The index sector covers 90 degrees of scanner rotation and is centered about the vertical reference line, leaving a 270-degree sector in which the scanner can receive signals. Signal pulses, shown in solid lines, occur when the beam lies between the tangent lines.

Angle θ_1, the angle between the vehicle vertical reference line and the line from the vehicle to the center of the planet cross section, is the angle which we wish to find. It establishes the position of a plane which contains the planet center and which is situated perpendicular to the scanning plane. Another such plane is established by the second scanning beam, and the intersection of the two planes so established is the local vertical line.

It is obvious that

$$\theta_1 = \tfrac{1}{2}(\tau_2 - \tau_1). \tag{1}$$

Angles τ_1 and τ_2 are measured quantities, and angle θ_1 is computed in accordance with (1).

A train of fixed-amplitude rectangular pulses is generated so that the duration of the pulses is equal to the time required for the scanning beam to sweep through angle τ_1. Such a train can be generated by gating a voltage on at the trailing edge of the index pulse and gating it off at the leading edge of the signal pulse. (The index pulse is established by magnetic pickoffs on the scanning mechanism.) It is evident that

$$\overline{V}_1 = E \cdot \frac{\tau_1^\circ}{360^\circ}, \tag{2}$$

$\overline{V}_1$ = average value of the pulse train, and
E = amplitude of the pulses.

Another train of pulses of amplitude E is generated, each pulse having a duration equal to the time between the trailing edge of the signal pulse and the leading edge of the index pulse. Its average value is

$$\overline{V}_2 = E \cdot \frac{\tau_2^\circ}{360^\circ} \cdot \tag{3}$$

Subtracting $\overline{V}_1$ from $\overline{V}_2$, we obtain a voltage proportional to θ_1. A similar process occurs in the signal channel associated with the other scanner, and a voltage is obtained which is proportional to an angle we will call θ_2, the angle defining the position of a second plane which contains the planet center.

It should be noted at this point that a voltage proportional to the angle subtended by the planet cross section can be derived by generating a similar train of pulses, each having a duration equal to that of the signal pulses. This angle will be used in computing vehicle altitude as shown in a later section.

Fig. 2 shows a sketch of the scanning assembly, in which two optical assemblies are driven at a continuous rate by the drive motor. In order to allow for vehicle spin rates of up to one revolution per second about an arbitrary axis, the optical assemblies are driven at 200 revolutions per second.

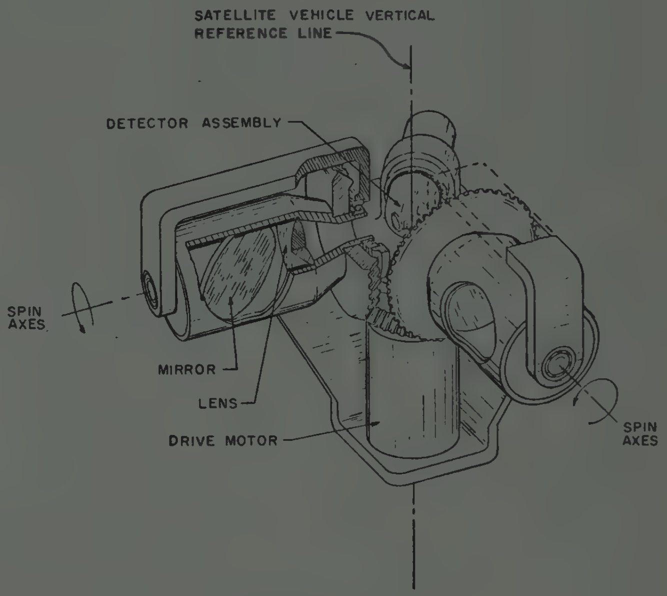


Fig. 2—Scanner assembly.

The optical system consists of a flat infrared-reflective mirror, a focussing lens, and a filter which attenuates radiation at wavelengths below 3 microns. It will serve to reduce direct solar radiation and minimize the effect of solar reflection gradients along the earth's surface. The focussing lens is a reflective doublet with an antireflection coating to give 94 per cent transmission at 4 microns.

Spectral distribution of radiation from the earth and atmosphere suggests a choice of operating wavelength greater than 3 microns. Because of the high scanning rate being used, the detector must have a relatively fast response time. In addition, it is desirable to use a detector which does not have to be cooled to liquid nitrogen temperature. Considering these requirements, an indium-antimonide detector (photoelectromagnetic type) was selected. A field of view of 2 degrees was chosen for the search beams. In order to achieve this condition with an optical system 1 inch in diameter and a focal length of $1\frac{1}{2}$ inches, a detector area of 1 square millimeter is required; a reasonable value.

Carrying a coolant for the detector is impractical for flights of significant duration, and another means of cooling is necessary. Outside the atmosphere, it is possible to cool the detector by radiation into space. An emissive surface thermally connected to the detector will be used to radiate into space, but since the sun will illuminate the vehicle at times, it is necessary to provide highly reflective shutters to cover the portion of the radiating surface directed toward the sun. The shutters must, of course, operate automatically.

VERTICAL TRACKING

In order to measure vehicle yaw, one axis of a single-degree-of-freedom gyro must be maintained parallel with the earth's vertical. When this is done, the orbital angular velocity provides a torque on the gyro rotor, causing it to seek a position perpendicular to the orbital plane. A synchro pickoff within the gyro provides a yaw angle signal. (Note that the gyro must have 360 degrees of freedom.)

Fig. 3 is used to describe the angles required to position the gyro. Angles θ_1 and θ_2 have been defined previously as those measured from the vehicle yaw axis to planes containing the earth center. Angles θ_1 and ρ are the ones required to properly orient the gyro. Line OE is assigned a value of unity to simplify the calculation, and the other line lengths are labelled in terms of θ_1, θ_2, and ρ. Eq. (4) is obtained by inspection, noting that $\tan\theta_2 = CA/OA$.

$$\tan\rho = \tan\theta_2\cdot\cos\theta_1. \tag{4}$$

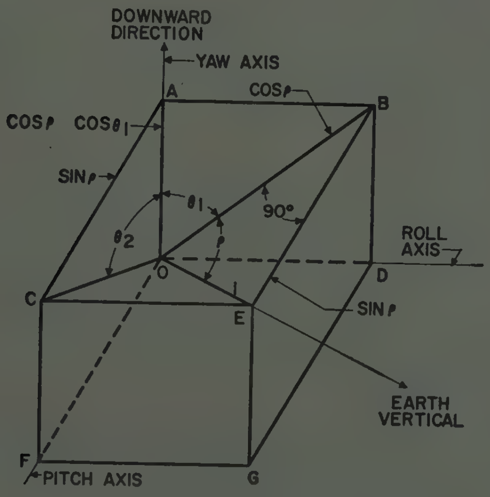


Fig. 3—Tracking angle relationships.

Voltages proportional to θ_1 and θ_2 are available from the detector channels and will be used to operate servo loops which maintain the gyro axis along the vertical. To accomplish this, a two-gimbal system is required with the gyro mounted on the inner gimbal. Fig. 4 shows a functional diagram of the servo loops.

Angle θ_1 and a conventional servo loop are used to position the outer gimbal. In addition to positioning the gimbal, the θ_1 servo positions a synchro for an output signal and operates a cosine potentiometer.

The dc signal proportional to θ_2 is altered by a shaping network to form $\tan\theta_2$, which is applied to the cosine potentiometer. The potentiometer output is a voltage proportional to the product $\tan\theta_2 \cos\theta_1$. Thus, the ρ servo generates a shaft position equal to ρ in accordance with (4). It drives a synchro to provide an output signal and positions the inner gimbal of the gimbal assembly.

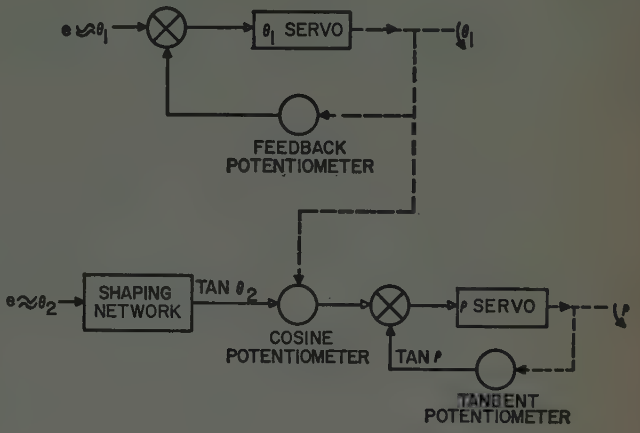


Fig. 4—Functional diagram of servo loops.

Use of a tangent feedback potentiometer in the servo loop at first appears impossible, since the tangent of 90 degrees is infinite; but when the value of either θ_1 or θ_2 is equal to or greater than 60 degrees, normal servo feedback loops are disconnected. A velocity memory loop is substituted so that the earth vertical is tracked even when there is no input from the scanners. The functional diagram of Fig. 5 shows the complete attitude reference system.

System response characteristics for yaw and pitch measurements are determined primarily by the servo loops. A tentative design of the servos was completed and the resulting loop for a representative case was simulated on an analog computer.

Required response time for both pitch and roll are assumed identical since the vehicle spin is taken to be in an arbitrary direction. Based upon that assumption, the roll servo presents the most difficult stability problem because of the nature of the feedback.

From an estimate of the weight and configuration of the load to be driven by the roll servo system, a load moment of inertia of 0.0025 slug-foot^2 was calculated. The motor selected to drive the load has the following parameters:

Rotor moment of inertia—	1.07 gm cm^2,
No-load speed—	6700 rpm, and
Stall torque—	0.63 oz-inch.

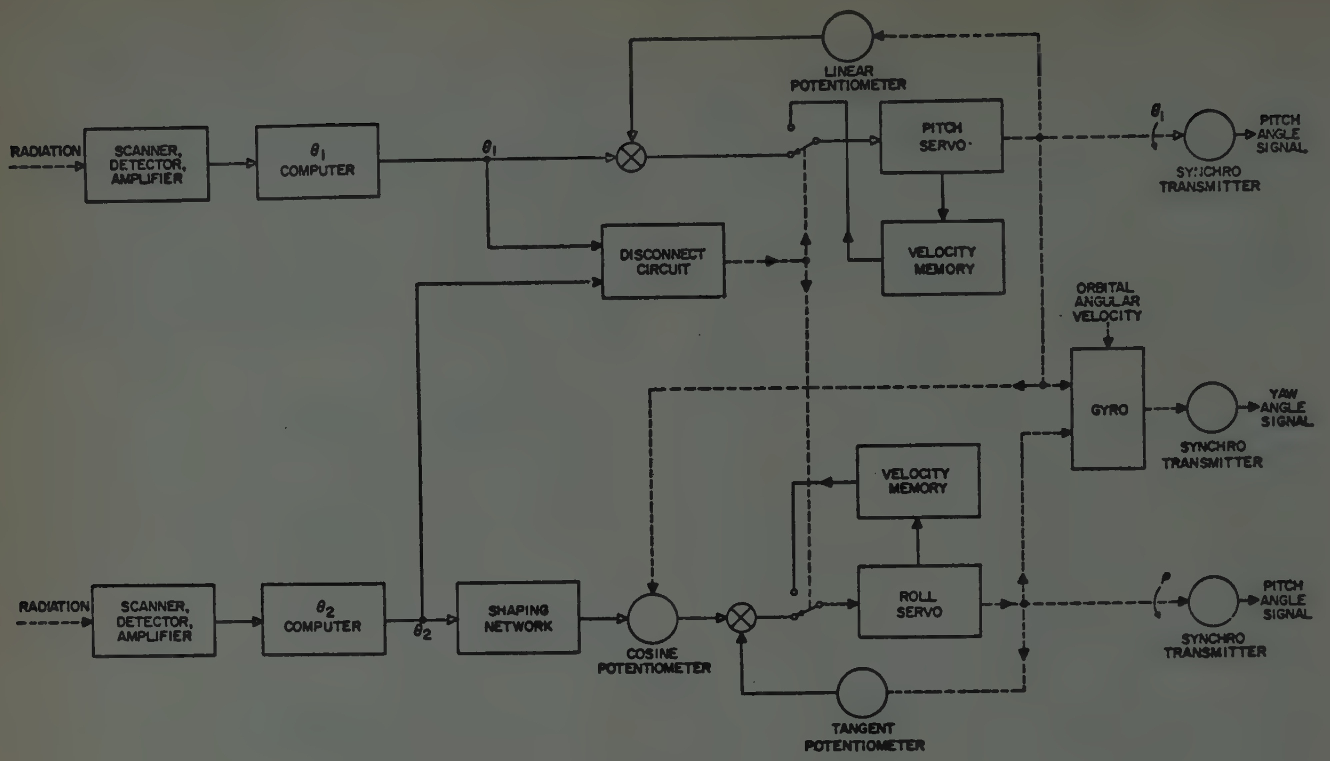

RADIATION
SCANNER, DETECTOR, AMPLIFIER
θ_1 COMPUTER
θ_1
LINEAR POTENTIOMETER
PITCH SERVO
θ_1
PITCH ANGLE SIGNAL
SYNCHRO TRANSMITTER
DISCONNECT CIRCUIT
VELOCITY MEMORY
ORBITAL ANGULAR VELOCITY
GYRO
YAW ANGLE SIGNAL
SYNCHRO TRANSMITTER
VELOCITY MEMORY
RADIATION
SCANNER, DETECTOR, AMPLIFIER
θ_2 COMPUTER
θ_2
SHAPING NETWORK
COSINE POTENTIOMETER
ROLL SERVO
PITCH ANGLE SIGNAL
SYNCHRO TRANSMITTER
TANGENT POTENTIOMETER

Fig. 5—Functional diagram of attitude reference system.

Maximum angular velocity of the load was taken to be 50 rpm, and a gear ratio of 80:1 was chosen, making the maximum operating speed of the motor 4000 rpm. With this gear ratio, the moment of inertia of the motor reflected to the load is 0.0005 slug-foot^2. Neglecting viscous friction other than that provided by the motor, the reflected damping term is 0.03 pound-foot second per radian. The resulting transfer function for the motor and gear train neglecting the effect of coulomb friction is

$$\frac{\rho}{E_m}(s) = \frac{k_t}{Nf_m(0.1s + 1)} \tag{5}$$

where

ρ = output angle, radians,
N = gear ratio = 80,
f_m = motor damping coefficient = 4.7×10^{-6} lb-foot second/radian,
k_t = motor torque constant, lb-foot/volt, and
E_m = motor control winding voltage, volts.

$$G(s) = \frac{\rho}{E_e}(s) = \frac{\mathrm{Kv}(0.5s + 1)(0.02s + 1)}{s(0.1s + 1)(5s + 1)(0.002s + 1)} \tag{6}$$

where $E_e(s)$ is the transform of the error voltage.

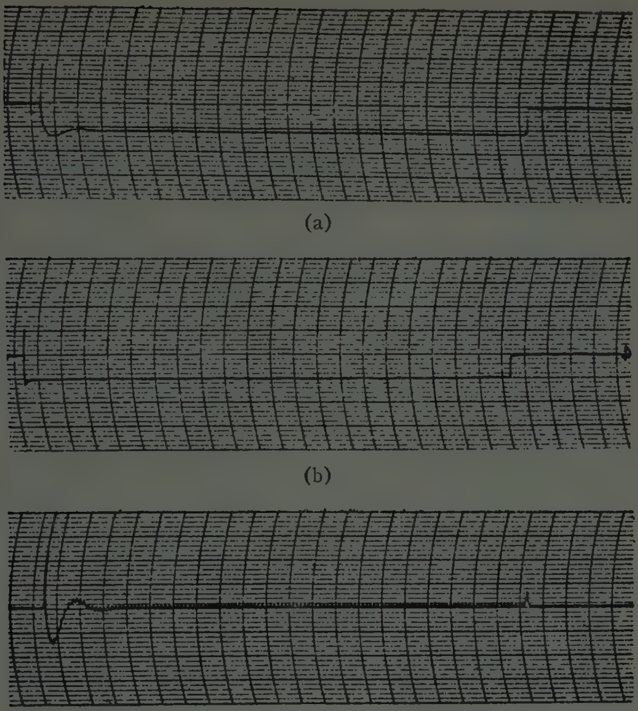
Closed-loop system response to a step input and to a ramp input is shown in Figs. 6 and 7, respectively. Kv was increased to 4800 for the recording in Fig. 8 to simulate the loop condition when ρ equals 60 degrees. The recording shows that the system is still stable.

ALTITUDE MEASUREMENT

In the system described here, the scanning planes do not, in general, intersect the earth center; hence the altitude cannot be calculated by using the subtended angle. Various methods for measuring altitude with the proposed system have been studied; the one which seems most desirable is described below. It should be noted that operation of the system relies upon satellite spin.

Angles θ_1 and θ_2 were described above. It is clear that if one of these angles, *e.g.*, θ_1, goes to zero, the scanning plane in which θ_2 lies must pass through the center of the earth. The magnitude of the angle subtended by the earth in the θ_2 plane at that time can be used to determine the vehicle altitude.

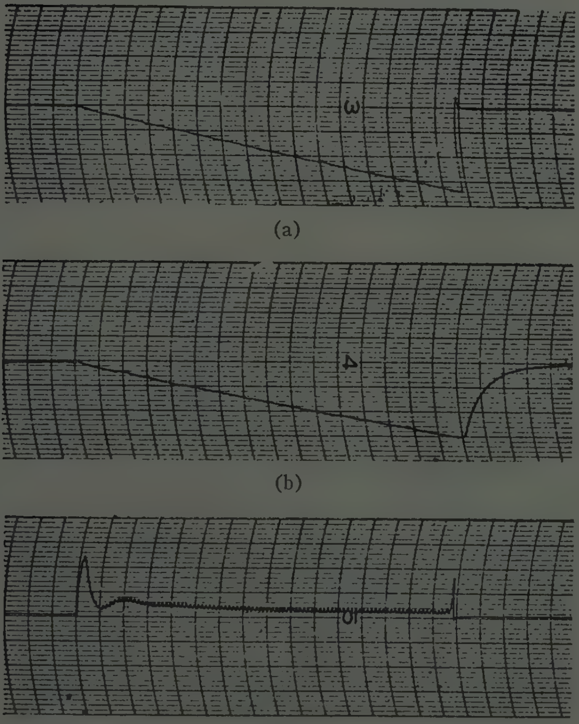
The angle subtended by the earth is measured by a sampling device when either of the scanning planes intersects the center of the earth. For example, the voltage proportional to θ_1 is applied to a voltage comparator which produces an output when θ_1 is equal to zero. The output voltage enables a sampler which applies a voltage β_2 to a memory unit as long as the enabling voltage is present. (β_2 is the angle subtended by the earth in the θ_2 scanning plane.) The memory unit output is used to compute altitude. Operation of the other channel is identical to that just described, so that β_1, the angle subtended by the earth in the θ_1 scanning plane, is applied to the memory unit when θ_2 is zero.


(a)

(b)

(c)

Fig. 6—Servo response to step input, ρ near 0°. Kv = 1200, paper speed = 50 mm/second unity feedback. (a) output—1 volt/mm, (b) input—1 volt/mm, (c) E_{ϵ_1} error voltage—0.2 volt/mm.

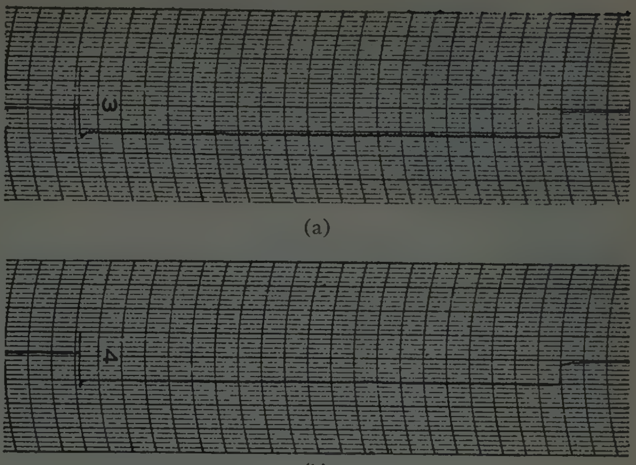


(a)

(b)

(c)

Fig. 7—Servo response to ramp input. Kv = 1200, paper speed = 50 mm/second, ramp voltage = 100 volts/second, unity feedback, (a) output—10 volts/mm, (b) input—10 volts/mm, (c) E_ϵ error—0.2 volt/mm.



(a)

(b)

Fig. 8—Servo response to step input, ρ near 60. Kv = 4800, paper speed = 50 mm/second, unity feedback. (a) output—1 volt/mm, (b) input—1 volt/mm.

Angle β is the output of the memory device, and is used to obtain altitude by means of the circuit shown in Fig. 9(b). Derivation of the relationship between altitude (h), angle β, and nominal earth radius (R), can be found by inspection of Fig. 9(a). Sin $\beta/2$ is formed by a diode shaping network, and the product $(h+R)$ sin $\beta/2$ is derived from the output of potentiometer P_1. This product is compared with a dc voltage proportional to R, and the servo output shaft position corresponds to $(h+R)$. A differential synchro with shaft input proportional to R provides the desired output, a signal proportional to h.

MISCELLANEOUS SOURCES OF ERROR

When the scanning system intercepts direct radiation from the sun, a considerable error may result. A fixed filter is employed to reduce the effect of solar radia-but the intensity of direct radiation will far exceed that received from the earth. Probably the major trouble which will arise from direct solar radiation will be momentary disabling of the detector. The detector will not be destroyed, but it may be disabled for a period up to 1 millisecond. A means of discrimination between the solar pulse and the rise in signal when the scanner goes from space to earth could be employed, based upon pulse duration. But if the detector is disabled for 1 millisecond, the scanner will sweep blindly through 72 degrees before recovering. There does not appear to be an easy way of avoiding this trouble, but it is not as serious as it seems at first glance. The percentage of time that the sun will appear in the scanning pattern will be small if the satellite is spinning. Even if a condition should exist in which the satellite motion is such that the scanning pattern is displaced very slowly, the probability that the sun will appear in the pattern is very small.

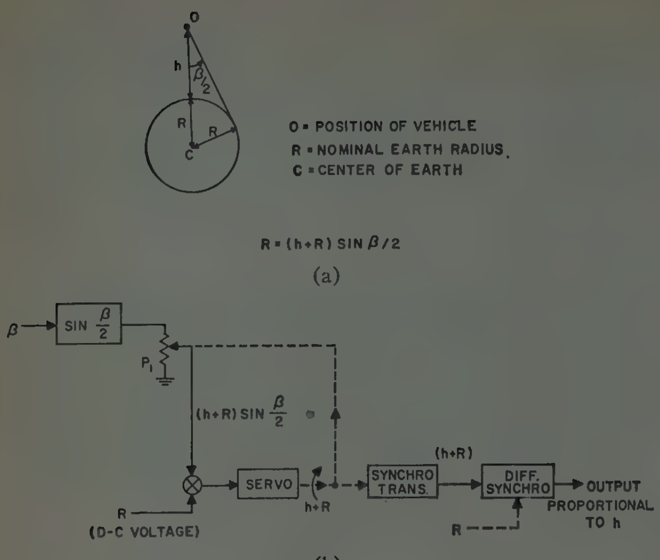


Fig. 9—(a) Derivation of expression for altitude. (b) Possible altitude computing circuit.

Errors will arise because of irregular terrain, atmospheric layers, and cloud formations; the use of a wider beamwidth may be necessary to reduce errors resulting from these effects. In addition, the effect of satellite spin upon the position of the scanning pattern will tend to average out the errors. These things will be only partially effective, and, if it becomes necessary, a nodding mirror might be employed to sweep the scanning planes and gain further error reduction.

In order to obtain an indication of the magnitudes of errors which might arise from natural sources, calculations were made for a two-dimensional case with perfect instrumentation assumed. A profile bulge of $0.005R$ was assumed to appear on one side of the body, and the other side was assumed to be perfectly smooth. The resulting error in location of the vertical and in computed altitude is given in Table I. Distances are given in units of planet radius.

In the above calculations a very narrow beamwidth has been assumed such that no error is encountered in determining when the beam is at the edge of the planet. When a finite beamwidth is used, some error will result. This effect has been studied for a conical beam shape, with the following assumptions:

TABLE I

Altitude	Vertical Error	Altitude Error
$0.02R$	0.76°	$-0.00263R$
$0.03R$	0.61°	$-0.00262R$
$0.04R$	0.52°	$-0.00264R$
$0.05R$	0.46°	$-0.00265R$
$0.06R$	0.42°	$-0.00267R$
$0.07R$	0.38°	$-0.00269R$
$0.08R$	0.36°	$-0.00271R$
$0.09R$	0.34°	$-0.00274R$
$0.1R$	0.32°	$-0.00278R$

1) It receives no energy from sources outside the beam and has uniform sensitivity within the field of view, and
2) radiation from the earth is uniform.

It is necessary to establish the instant at which the beam center line, chosen as a reference line, is tangent to the earth. Two alternative operations on the signal were considered:

1) Differentiate the signal and let the maximum of the resulting signal denote the point of tangency, or
2) select a particular ratio of signal level to maximum signal level to represent that instant.

The first method was chosen because it appears to be more practical from the standpoint of instrumentation. Error in altitude caused by the error in determining true reference line tangency by this method is given by the expression

$$\frac{\Delta h}{h} \cong \frac{(a+1)(a+2)\alpha^2}{2 - a(a+2)\alpha^2} \tag{7}$$

where

Δh = altitude error,
h = true altitude,
α = one-half of the apex angle of the beam, and
$a = h/R$, ratio of altitude to earth radius.

It should be noted that the variation of error is known as a function of altitude; hence, it is theoretically possible to compensate for it. It is doubtful, however, that such compensation will need to be used since a value as large as 0.1 for α results in errors which are small compared to those caused by terrain irregularity.

High-Speed Electrometers for Rocket and Satellite Experiments*

J. PRAGLIN†, MEMBER, IRE, AND W. A. NICHOLS‡, MEMBER, IRE

Summary—**Highly stable micromicroammeters capable of measuring currents as low as 5×10^{-14} amperes with a frequency response of at least 30 cps can be realized using directly-coupled electrometer circuitry. The circuit can be packaged in approximately 9 cubic inches and requires less than 100 mw of power. An analysis of the speed of response and signal-to-noise ratio is given in the text and expressions are derived to show the theoretical possibilities of the method and to serve as design guides for realization of the maximum possible performance. Several practical circuits are shown and a summary is presented of the number of circuits which have been used in rocket and satellite exploration.**

INTRODUCTION

TRANSDUCERS such as ion chambers, photomultiplier tubes, proportional counters and ionization gauges are used to gather much of the experimental data obtained on rocket and satellite probes of space and the upper atmosphere. In the range of measurement usually encountered, the output of these devices is restricted to currents in the order of 10^{-7} to 10^{-13} amperes. Considerable amplification is then required to operate the input of a telemetering system. In general, this paper will deal with the theoretical and practical background needed to design suitable current amplifiers for this application and will describe in some detail circuits designed by the authors for the measurement of ultraviolet light and soft X-ray radiation from the sun and other celestial bodies. The circuits measure the output of a gas filled ion chamber and have a 5×10^{-12}-ampere full scale sensitivity, a frequency response of about 20 cps and a zero stability of better than 10^{-13} amperes per week.

AVAILABLE METHODS

A survey of the available types of voltage and current amplifiers will immediately narrow down the possible devices to either a dc vacuum tube electrometer amplifier or an ac amplifier used in conjunction with a capacitor modulator. The vibrating reed capacitor modulator can be immediately eliminated because of the size and power requirement of available units, the complexity of the necessary circuitry, and the difficulties inherent in modulator circuitry when speed of response is to be obtained with large feedback factors. Semiconductor devices used either directly-coupled or as modulators, or mechanical modulators do not possess either the current stability or the low noise necessary for measurement over most of the desired current range.

In keeping with the above then, it is required that the input be an electrometer vacuum tube, and specifically, because of size and performance considerations, the type CK5886, electrometer tube manufactured by Raytheon. This tube possesses the following characteristics: The input or grid current is typically less than 2×10^{-14} amperes. The input resistance is greater than 10^{14} ohms. The voltage stability with regulated plate and filament supplies is typically better than plus or minus 50 mv per week. The input current noise is typically less than 2×10^{-15} amperes peak to peak and the short circuit input noise is less than 100 μv peak to peak. The tube is a filamentary type which requires 12-mw filament power and can be operated satisfactorily with as little as 10 volts of plate supply potential.

STABILITY AND SPEED OF RESPONSE PROBLEM

The simplest method of measuring these currents with an electrometer tube is shown in Fig. 1. The tube is arranged as a voltmeter and the drop across a known value of resistance is used as a measure of the current. One difficulty then immediately becomes apparent. It has been stated above that the electrometer tube can be relied on to have a 50-mv voltage stability. It is therefore necessary that the current signal be considerably greater than this value if the device is to have negligible zero drift. Then, for example, if we desire an instrument to measure 5×10^{-12} amperes with less than 1 per cent drift per week, it is necessary to use a 10^{12}-ohm grid resistor in order that 5×10^{-12} amperes generate a 5-volt signal. Resistances of this magnitude are entirely useable and stable measurements can be made. However, the method suffers two severe limitations. First, if any amount of capacitance is present in the in-

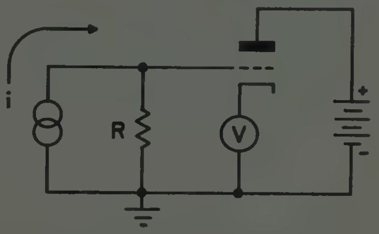


Fig. 1—Simple micromicroammeter employing voltage drop across a shunt resistor.

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* Original manuscript received by the IRE, November 23, 1959. The work reported here was performed at Keithley Instruments, Inc., Cleveland, Ohio. Parts of the work were performed under contracts to the Navy, Space Agency, and the Air Force.

† Keithley Instruments, Inc., Cleveland, Ohio.

‡ Naval Research Lab., Washington, D. C.

put circuitry, the response speed can be extremely slow since, for example, a 10^{12}-ohm resistor in parallel with a 10-mmf capacitor gives a 10-second time constant. This response speed would be prohibitively slow in many rocket applications. Second, a less severe limitation is that the ammeter drop is 5 volts. By the proper use of negative feedback, however, this dilemma can be resolved and fast, stable micromicroammeters can be constructed whose speed of response is relatively unaffected by input capacitance.

The method first used was first described in some detail by Pelchowitch and Zaalberg Van Zelst[1] and their article should be consulted for another equally valid analysis of the transient response problem. The basic circuit employed is shown in Fig. 2. A phase-inverting amplifier, A, is connected as shown. C_1 represents the sum of the tube input capacity and cable capacity; i_{in} is the input current; e_0 is the output voltage; C_2 is either the stray capacity or a lumped capacity placed across resistor R_1, the current-measuring resistor. Neglecting, for the moment, C_1 and C_2, to arrive at the dc response, the current i_{in} flows into the input terminal and through shunt resistor R_1. Since A is an amplifier of gain k,

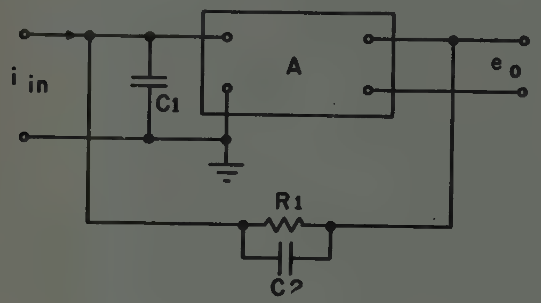

Fig. 2—Feedback micromicroammeter circuit.

$$e_{\text{in}} = e_0/k. \tag{1}$$

Since all the current i_{in} must flow through R,

$$i_{\text{in}} = -\frac{e_0}{R_1} + \frac{e_0}{R_1 k}. \tag{2}$$

Now if k, the loop gain is large, i_{in} is simply e_0/R_1. The effective input resistance is simply $e_{\text{in}}/i_{\text{in}}$ and from (1) and (2) the effective input resistance, R_{in} is given by

$$R_{\text{in}} = \frac{R_1}{k+1}. \tag{3}$$

Therefore, the use of a negative feedback circuit as shown in (1) reduces input drop by a factor equal to loop gain and consequently by (3) gives the ammeter a lower effective input resistance. Also by (2), if the loop gain is large, the current sensitivity is precisely the same as if the current were being measured by means of a voltmeter and shunt resistor in Fig. 1. Then, the zero drift of the circuit will be precisely the same as in the voltmeter case provided that the circuit is being used with a current source, as is always the case in these applications.

From (3) it can be gathered that the speed of response should be increased since the effective input resistance of the ammeter is reduced. This will be shown more exactly below. In any case, however, if a substantial increase in response speed is to be obtained, the gain, k, must be large and to this end feedback factors of as much as 10^5 can be used. Various methods can be used to stabilize the loop against oscillation. One of the authors[2] has described a method to be used where it is desired to achieve stable response with R_1 at any value from zero to 10^{12} ohms. This system employs stabilizing networks internal to the amplifier to prevent oscillation. However, this method does not permit optimum performance at very low currents. The other approach is simply to let the lag formed by the input capacity C_1 and the high megohm resistor R_1 be dominant and stabilize the loop. The time constant of this lag will be sufficient in most cases to stabilize loop gains as high as 10^5. For example, with 100-$\mu\mu$f input capacitance and a 10^{12}-ohm measuring resistor this network forms a corner at 0.0016 cps. If the value of R_1 is reduced to the point where this corner is not sufficient, the loop gain may be reduced or the method discussed by Praglin[2] may be used.

ANALYSIS OF THE RESPONSE SPEED

For the purposes of circuit analysis, Fig. 2 is modified as shown in Fig. 3 to include a network in the feedback loop designated by B, the transfer function of the network. The assumption is made that k is real over the bandwidth of interest. Then the equivalent circuit is as shown in Fig. 4. The input current is given by

$$i_{\text{in}} = e_{\text{in}} p C_1 + [e_{\text{in}} - \beta e_0] \frac{1 + pC_2R_1}{R_1} \tag{4}$$

where p is the differential operator. Rearrangement of terms yields:

$$i_{\text{in}} = -e_0 \left[\frac{pC_1}{k} + \frac{1 + pC_2R_1}{kR_1} + \frac{B(1 + pC_2R_1)}{R_1} \right]. \tag{5}$$

Finally, if the circuit is excited by a current square-wave, $i_{\text{in}}[U(t)]$, where $U(t)$ is the Heaviside step function and (5) multiplied by this function, the solution of the equation can easily be obtained by means of the Laplace Transformation to give

$$e_0 = \frac{-i_{\text{in}} R_1 k}{1 + Bk} \left[1 - \exp \left(\frac{-t}{R_1 \left[\dfrac{C_1}{1 + Bk} + C_2 \right]} \right) \right]. \tag{6}$$

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[1] I. Pelchowitch and J. J. Zaalberg Van Zelst, "A wide-band electrometer amplifier," *Rev. Sci. Instr.*, vol. 23, pp. 73–75; February, 1952.

[2] J. Praglin, "A new high stability micromicroammeter," IRE TRANS. ON INSTRUMENTATION, vol. I-6, pp. 144–147; June, 1957.

The time constant, T, is then

$$T = R_1\left[\frac{C_1}{1+Bk} + C_2\right] \tag{7}$$

where it is seen that the time constant formed by the input capacitance RC_1 is reduced by the feedback factor Bk. On the other hand, RC_2, the time constant formed by the resistor and any capacity across it is not affected. This fact may be used to damp the response of the amplifier by adding capacity across C_2 if desired, but in high-speed circuits this time constant must be neutralized if any speed of response is to be obtained. To this end let the network in the box labeled B in Fig. 3 have the form of the network in Fig. 5. The transfer function of this network is

$$B = \frac{1}{1 + pC_3R_2}\,. \tag{8}$$

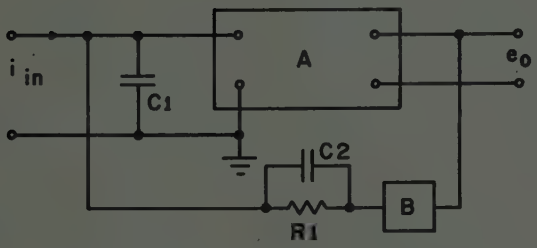


Fig. 3—Feedback micromicroammeter employing a correcting network in the feedback loop.

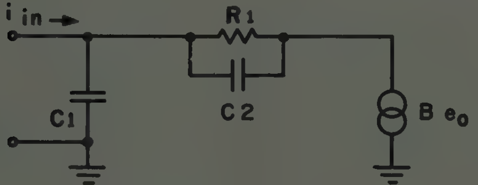


Fig. 4—Equivalent circuit of Fig. 3.

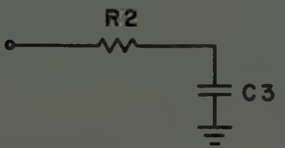


Fig. 5—Correcting network for feedback loop.

If this transfer function is substituted for B in (5) and if the condition is met that $C_3R_2 = C_2R_1$, then the third term in the parenthesis is simply $1/R_1$. Now if the Laplace Transform is used to give the network response to current step $i_{\text{in}}[U(t)]$, the result is as follows:

$$e_0 = \frac{-i_{\text{in}}R_1k}{k+1}\left[1 - \exp\left\{\frac{-t}{R_1\left[\dfrac{C_1+C_2}{k+1}\right]}\right\}\right] \tag{9}$$

and the time constant is

$$T = \frac{R_1(C_1+C_2)}{k+1} \tag{10}$$

which now shows that if the above lead and lag networks are equal all time constants associated with the circuit are degenerated by loop gain.

This analysis ignores the fact that if a resistor of 10^{11} or 10^{12} ohms is suspended near a ground surface it is inevitable that stray capacity will be distributed all along the resistance element and the performance will resemble that of a delay line. It is therefore impossible to perfectly compensate the circuit so that the high megohm resistor, R_1, looks entirely resistive. In general, this consideration and the consideration of signal-to-noise ratio to be discussed presently limits the attainable speed of response. It is the usual experience that with a 10^{12} ohm resistor and very little input capacity it is possible to achieve a response of approximately 50 cps.

SIGNAL-TO-NOISE RATIO

The noise at the input of the micromicroammeter may be contributed by three sources:

1) Internal noise generators in the electrometer tube.
2) Grid current noise.
3) Johnson noise.

Internal noise generated in the electrometer tube can be represented in Fig. 6 as a voltage generator, e_n, in the cathode circuit. Grid current noise may be represented as an additional current generator, i_g, equal to $\overline{i_g^2(f)_{\text{av}}}^{1/2}$ as can the Johnson noise by a generator, i_j equal to $\overline{4ktf/R_1}^{1/2}$. Thus the total current flowing into the input equals $i_{\text{in}}+i_j+i_g$. Neglecting e_n for the moment, this noise current is dependent on frequency exactly as is the input current. For this analysis the steady-state response to sinusoidal stimuli is sufficient, and if we again employ (5), but substitute $j\omega$ for p and (8) for B, we obtain

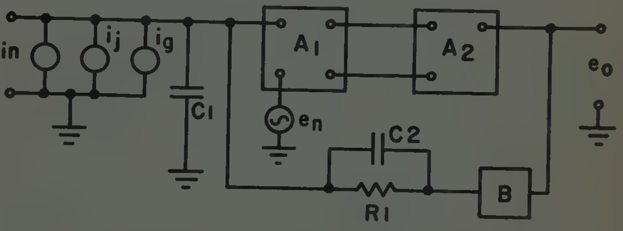


Fig. 6—Micromicroammeter with noise generators.

$$e_0 = \frac{-i_{\text{in}}R_1k}{R_1(C_1+C_2)}\left[\frac{1}{j\omega + \dfrac{k+1}{(C_1+C_2)R_1}}\right] \tag{11}$$

with a pole at $-[k+1/(C_1+C_2)R_1]$. Evaluation along the $j\omega$ axis shows that e_0 goes to zero with increasing frequency. Since i_{in} in (11) can apply to i_j, i_g, i_{in} or their sum shows that these noise sources go to zero with frequency exactly as does the signal or that at any frequency these noise sources and the input current form an unchanging ratio. Specifically, the amplitude of e_0 is given by

$$e_0 = \frac{-i_T R_1 k}{(C_1 + C_2)R_1}\left[\frac{1}{\omega^2 + \left\{\dfrac{k+1}{(C_1 + C_2)R_1}\right\}^2}\right]^{1/2} \quad (12)$$

where i_T represents either the sum of i_{in}, i_j, and i_g or any one taken separately.

In general, in an electrometer circuit of this type the terms i_g and i_j are nearly negligible since the output signal is generally never less than 1 volt. The grid current of a 5886 tube is less than 2×10^{-14} amperes, measured as dc quantity, and fluctuation of this quantity about the mean dc value, which we may regard as the peak-to-peak ac grid current noise, usually does not exceed 2×10^{-15} amperes, and of course, the rms value is considerably smaller. So that, if 10^{-12} amperes of current is to be measured, the grid current noise is not more than 0.2 per cent of the signal peak-to-peak. The Johnson noise expressed as an equivalent current noise from the above expression, an assumed bandwidth of 50 cps, room temperature, and a 10^{12}-ohm resistor, amounts to about 10^{-15} amperes rms or about 5×10^{-15} amperes peak-to-peak. If smaller resistors are used, provided that the current to be measured is increased proportionally, the signal-to-noise ratio with respect to Johnson noise will be even more favorable.

The effect of internal noise sources in the electrometer tube is markedly different since this noise source is not attenuated by the circuit capacities. A brief qualitative description of the effect is helpful here. Consider the circuit of Fig. 3 with C_2 compensated for. As the frequency increases, the impedance of C_1 falls and as a result the signal obtained at the amplifier input decreases. At the same time it is the ratio

$$\frac{j\omega C_1}{R_1 + j\omega C_1}$$

which determines the feedback factor. Thus, automatically, as the signal falls at the amplifier input, the gain is increased in proportion. The result is that, as has been shown analytically, the apparent time constant of the input circuit is decreased by loop gain. Actually the physical effect of C_1 is in no way changed. We merely compensate for the attenuation caused by C_1 by boosting the forward gain of the amplifier. The difficulty with this arrangement is that the forward gain of the amplifier cannot increase beyond some practical limit since the internal noise of the first stage is constant over the entire band and is increasingly amplified as the forward gain of the amplifier increases to compensate for the capacity loading of C_1. Therefore, if the average peak-to-peak noise of an electrometer tube is 100 μv and the loop gain of the amplifier is 10^4, at some frequency the noise in the output will amount to 1 volt peak-to-peak.

A more sophisticated analysis is as follows. Let us consider the noise source e_n of Fig. 6; e_n represents the internal tube noise. Other sources of noise will be regarded as minor contributors in light of the above discussion, and their consideration will be omitted. Now from Fig. 6.

$$e_0 = k\left[e_n - \frac{e_0 B}{1 + \dfrac{j\omega C_1 R_1}{1 + j\omega C_2 R_1}}\right]. \quad (13)$$

Now if, as before, B is picked to neutralize the lead formed by R_1C_2, (13) becomes

$$e_0\left[1 + \frac{k}{1 + j\omega(C_1 + C_2)R_1}\right] = ke_n \quad (14)$$

and is finally reduced to the form

$$e_0 = ke_n\left[\frac{j\omega + \dfrac{1}{(C_1 + C_2)R_1}}{j\omega + \dfrac{1+k}{(C_1 + C_2)R_1}}\right]. \quad (15)$$

If we are at midband, the dc signal is given by $e_0 = i_{\text{in}}R_1$, and, therefore, the signal-to-noise ratio, S, is given by

$$S = \frac{i_{\text{in}}R_1}{ke_n}\left[\frac{j\omega + \dfrac{1+k}{(C_1 + C_2)R_1}}{j\omega + \dfrac{1}{(C_1 + C_2)R_1}}\right], \quad (16)$$

where e_n is interpreted as the peak-to-peak noise. By the rules of complex algebra, the amplitude is given by

$$S = \frac{i_{\text{in}}R_1}{ke_n}\left[\frac{\omega^2 + \left[\dfrac{1+k}{(C_1 + C_2)R_1}\right]^2}{\omega^2 + \left[\dfrac{1}{(C_1 + C_2)R_1}\right]^2}\right]^{1/2}. \quad (17)$$

When ω is large, (17) reduces to

$$S = \frac{i_{\text{in}}R_1}{ke_n} \quad (18)$$

which says that the signal-to-noise ratio will eventually deteriorate to $i_{\text{in}}R_1/ke_n$. It is, therefore, of no use to make the loop gain exceed

$$k = \frac{i_{\text{in}}R_1}{S'e_n}, \quad (19)$$

where S' is the tolerable signal to noise ratio.

Eq. (17) also shows that the signal-to-noise ratio is markedly affected by the size of C_1, the input capacitance. (In most cases it will be a valid assumption to consider C_2 negligible in comparison to C_1.) If we consider the bracketed term on the right side of (17) it can be seen that if C_1 is very small,

$$\frac{1+k}{(C_1+C_2)R_1} \quad \text{and} \quad \frac{1}{(C_1+C_2)R_1}$$

are large quantities and ω must be correspondingly large before the signal-to-noise ratio deteriorates. If C_1 is a large value, it allows the ω terms to predominate at a much lower frequency. Therefore, we may conclude that the signal-to-noise ratio must deteriorate progressively with frequency; and there is a maximum loop gain which is of use as expressed by (19) for a maximum tolerable signal-to-noise ratio. Secondly, the value of C_1, the input capacitance, largely determines the signal-to-noise ratio for a specified gain and frequency band as stated in (16). Thus, since (19) specifies that the signal-to-noise ratio will reach in the limit a value dependent only on the input signal, the measuring resistor, the gain k, and the first stage tube noise, it is a waste of resources to increase the noise by increasing the loop gain beyond the k specified in (19). Now with this value of k the maximum allowable value of C_1+C_2 for a given rise time is given by (9) or by (12) for a given frequency response.

Obviously, by (17) at frequencies below those at which the bracketed term on the right is equal essentially to 1, the signal-to-noise ratio will be better and it may be argued that an improvement in signal-to-noise ratio will result with filter at the output of the amplifier to cut off the response at some lower frequency. In theory this is not true because this is in no way different from reducing loop gain. As shown by (12) the amplifier frequency response falls off on reduction of k precisely as if a low-pass filter were placed at the output. There are, nevertheless, several advantages to be gained from an output filter in practice. This is so because the input tube is usually a filamentary type tube with a pronounced mechanical resonance at about 5000 cps. Usually this frequency is far higher than the maximum operating frequency and it is of advantage to eliminate this frequency and any other sources of extraneous noise from the output.

LOGARITHMIC CIRCUITS

The resistor R_1 may be replaced by a logarithmic element such as a thermionic diode operated in a space charge limited region or a silicon diode operated on a limited portion of its forward biased region. In general, the authors have preferred the use of a 5886 electrometer tube, diode connected, as a logarithmic element, due to the wider range of measurement available (approximately 10^{-5} to 10^{-13} amperes in one range vs a range of about 10^{-4} to 10^{-10} amperes for silicon diodes), much less temperature dependence and far greater uniformity of characteristic between units than silicon offers.

Unfortunately, with logarithmic circuits, it is not possible to employ the compensation schemes used for resistive feedback. This is so since the voltage current relationship for a thermionic logarithmic current element is

$$V = A \log i_{\text{in}} + B, \tag{20}$$

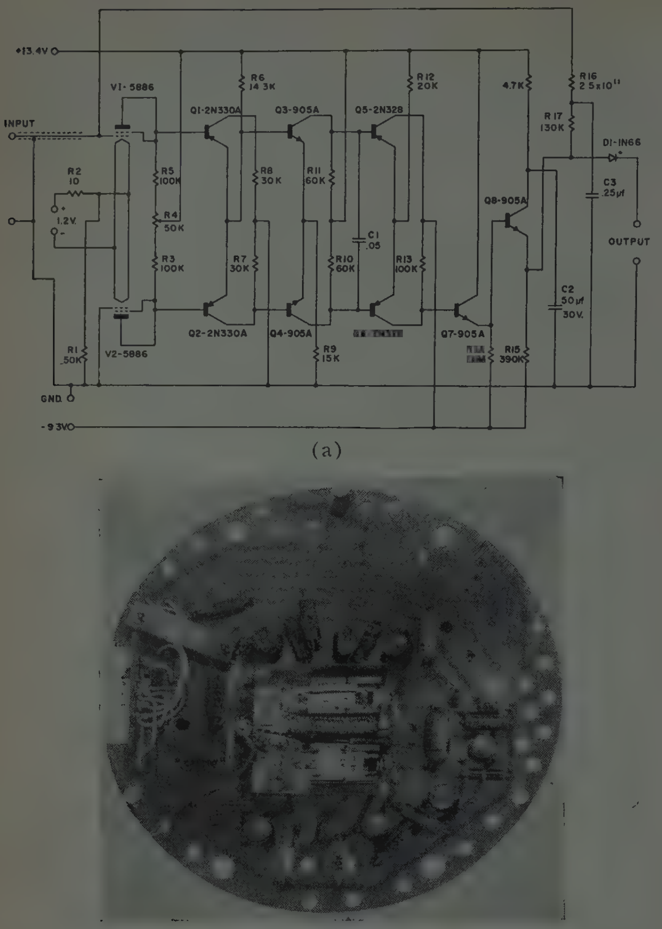
where A is the slope of the voltage current relationship and B is the residual dc voltage in the thermionic log element due to the initial emission velocity of electrons. Differentiating both sides of (20) we obtain

$$\frac{dV}{di} \equiv \frac{\Delta V}{\Delta i} = \Delta R = \frac{a}{i}, \tag{21}$$

where ΔR represents the differential resistance at some current i. Here, therefore, the feedback resistance, if the diode is used in a feedback loop, is different for different values of input current so that it is impossible to compensate for the feedback capacitance inevitably present in the diode except over a very small range. In a log circuit with a several decade range, if the circuit is critically compensated at the low-current end, the circuit will be hopelessly over-compensated at the high-current end. It is then necessary that a log circuit be overdamped at the low-current end and just achieve critical damping at the high-current end if reasonable transient response is to be achieved. Also, since the effective feedback resistance is always changing it usually is not possible to stabilize the circuit by letting the input time constant act as a dominant lag network, but additional attenuation must be supplied within the loop. If the transient response of a log circuit is considered on a small signal basis, each time sampling a different portion of the range, linear theory can apply and the method referred to by Praglin[2] can be used to analyze the results. On a practical basis, a frequency response of about 1 cps can be achieved at a current of 10^{-12} amperes, a response considerably poorer than that of a linear circuit. Logarithmic circuits, however, have proven extremely useful in many types of exploratory rocket and satellite experiments where either the range of measurement was in doubt or an extremely wide range of results were expected.

PRACTICAL CIRCUITS

The principal requirements outlined in the theoretical section require that the input circuit consist of an electrometer tube followed by sufficient gain for the desired performance. The first type of circuit used by the authors is shown in Fig. 7(a). In this circuit two electrometer tubes were employed in a balanced configuration to cancel expected changes in filament voltages. These tubes are followed by a 5-stage transistor amplifier. This circuit and modifications of it have been used in numerous NRL rocket flights and a version of this circuit is in use in the Vanguard 20-inch satellite. The maximum frequency response has been about 30 cps for a sensitivity of 5×10^{-12} amperes. For less sensitive


+13.4V
INPUT
V1-5886
R6 14.3K
R12 20K
4.7K
R16 2.5x10¹¹
Q1-2N330A
Q3-905A
Q5-2N328
R17 130K
D1-1N66
R2 10
R5 100K
R8 30K
R11 60K
C3 25μf
Q8-905A
1.2V
R4 50K
C1 .05
OUTPUT
R3 100K
R7 30K
R10 60K
R13 100K
C2 50μf 30V
Q2-2N330A
Q4-905A
Q7-905A
R9 15K
R15 390K
R1 50K
V2-5886
GND
-9.3V

(a)

(b)

Fig. 7—Vanguard Satellite Photograph. Micromicroammeter. (a) Circuit diagram. (b) The electrometer tubes are shown in the center and are normally covered with a shield. The extra space at the top of the picture has been provided for a transistor converter power supply.

circuits the frequency response is correspondingly greater. The zero stability is in the order of about plus or minus 2 per cent of full scale per week. The power consumption of the circuit is in the order of 100 mw and the output is arranged to supply 5 volts at full scale to a telemetering system. In general, the performance has been satisfactory; however, one major difficulty has been experienced. In this configuration considerable noise contribution is obtained from the first transistor stage. This noise has the typical $1/f$ semiconductor noise distribution. This semiconductor noise is particularly annoying since it predominates at low frequencies where it is desired to make measurements. In general, tube noise except for the higher frequency mechanical resonances of the tube structure did not seem to display any noticeable $1/f$ distribution. It was possible to select germanium or silicon transistors which had satisfactory noise for the application. However, this procedure was both time consuming and annoying since a relatively small percentage of transistors, particularly silicon, would give satisfactory performance; and it was usually necessary to use silicon transistors because of the usual temperature considerations.

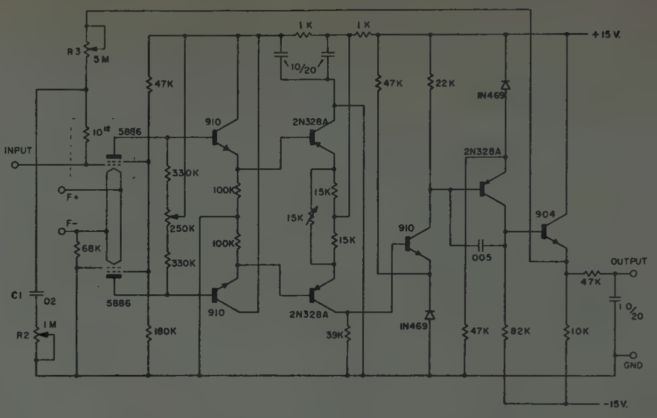
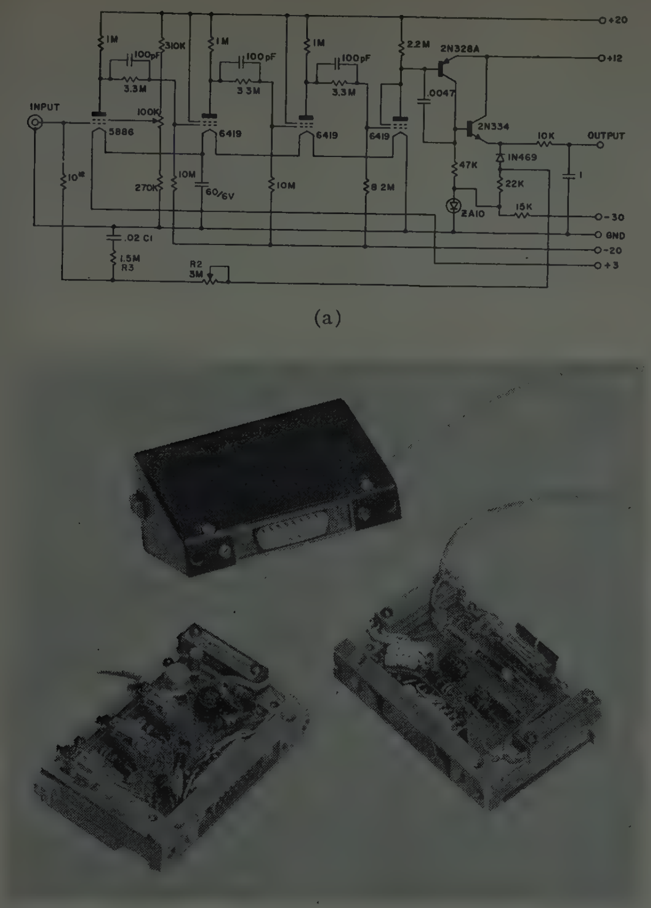

R3
5M
47K
1K
1K
10/20
+15V
47K
22K
1N469
10¹² 5886
910
2N328A
INPUT
330K
100K
2N328A
F+
15K
15K
F-
250K
100K
15K
910
904
68K
330K
.005
OUTPUT
47K
C1
.02
5886
910
2N328A
1N469
10/20
R2
1M
180K
39K
47K
82K
10K
GND
-15V

Fig. 8—Circuit schematic diagram of improved balanced input electrometer tube and transistor circuit.

An alternate circuit which offered some improvement over the first circuit is shown in Fig. 8. Balanced electrometer tubes are still used but the electrometer tubes are followed by emitter-follower stages. With silicon transistors it is possible to achieve about a 1-meghohm plate circuit impedance and if the 5886 tubes are then connected as pentodes it is possible to obtain a voltage gain of about 10. In the previous circuit the voltage gain from the grid to plate of the electrometer tubes is less than 1.0. This arrangement, as might be expected, offers a 10:1 reduction in transistor noise over the first circuit through a better impedance match between the electrometer tubes and the transistor stages. With this circuit, transistor selection is not as critical although some selection is still necessary. As far as other performance specifications and power consumption, it is essentially the same as the first circuit.

The requirements of low power consumption had, at the beginning of the project, seemed to dictate the use of transistors wherever possible. Use of a balanced input stage had been made necessary by the possibility of shifting battery voltages. However, due to the difficulties experienced with semiconductors, the possibility that a tube amplifier would give better results seemed worth considering. Also the requirements for operation from batteries had changed to operation with transistor inverter supplies. With these supplies it is not much more difficult to provide very close regulation of output voltages with zener diodes and transistorized feedback regulators. The circuit in Fig. 9(a) shows the new approach. Four subminiature tubes are used in the main amplifier and two transistors in the output stage. The power requirement for the series filament string is only 3 volts at 10 ma or 30 mw. Since it is possible to run subminiature tubes at far lower plate currents with adequate gain than it is possible to operate transistor collector circuits, the total plate current requirement is 40 μa at 20 volts or 0.8 mw. The output transistor stages dissipate about 8 mw at no signal. The total power requirement is then less than 40 mw against perhaps 100 mw for the previous circuit. The circuit is



(a)

(b)

Fig. 9—(a) Circuit schematic diagram of subminiature tube circuit employing transistors only in the output circuit. (b) Photograph of the circuit of (a). The dimensions are 2×3×1½ for the largest part of the wedge cross section. At the top is the complete unit with the case. Below, at the left, the side carrying the three 6419 tubes and the output transistors is shown. At the right the input stage with the electrometer tube and the high megohm resistor is seen.

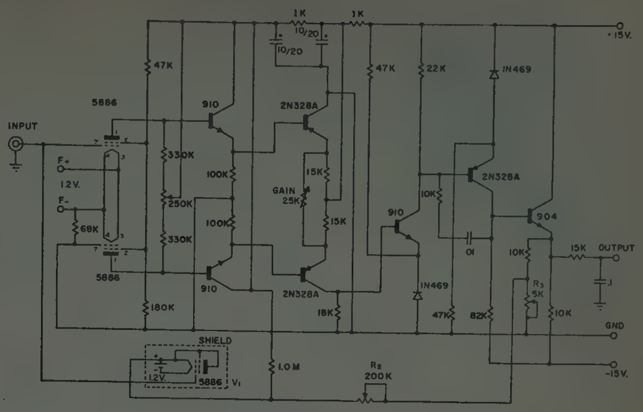


Fig. 10—Circuit schematic diagram of a logarithmic micromicroammeter using type 5886 tube connected as a diode.

characterized by a signal-to-noise ratio with no component selection which is consistently at least as good as the best performance of the previous two circuits with selected transistors.

In all three circuits a network, which has been referred to (Fig. 3) in the Theoretical Section, has been employed to compensate the response. In early circuits, the exact form of compensation circuit referred to in the article was used, and fixed value components were soldered in. However, in the circuits of Figs. 8 and 9 the compensation network is adjustable by means of carbon potentiometers for greater convenience. It was also found that, in practice, the trimming of the circuit was facilitated by adding an additional resistance in series with the capacitor of the compensating network and also making this additional resistance variable. (See R_3, C_1, and R_2 of Figs. 8 and 9). This additional resistance helps to compensate for excess phase shift introduced by the distributed nature of the stray capacitance to the high meghohm resistor body and the phase shift introduced in some cases by the amplifier. Figs. 7(b) and 9(b) are photographs of two of the circuits.

In Fig. 10 is shown a logarithmic circuit. The general approach is similar to the linear circuits except for the log diode. The power supply for the logarithmic element, V_1, is floating and potentiometer R_2 taps off a voltage to buck-out the initial emission voltage of the diode. This particular circuit has a range of 10^{-5} to 10^{-11} amperes without scale changing. Potentiometer R_3 determines the forward gain of the amplifier and hence the slope of the log voltage current relationship. This circuit has a frequency response of about 7 cps at 10^{-11} and several thousand cps at 10^{-5} amperes. The stability after a short warmup is better than 5 per cent of a decade per day.

METHODS OF TESTING TRANSIENT RESPONSE

It is extremely difficult, by ordinary methods, to generate current step functions or sinusoidal waves of known amplitude in the region of 10^{-8} to 10^{-12} amperes. The use of a resistor with a known potential is useless except at dc because of the stray capacities involved. In the course of this investigation two methods have been found satisfactory.

1) The generation of current wave forms by the modulation of light input to a photocell.

2) The use of a triangular wave generator with a small capacitance to generate a current step function.

The first method is free of any theoreitcal objections and lends itself, with a little ingenuity, to the generation of both sine and step functions of current. The main drawback is that it is difficult to find phototubes with a low enough dark current to make the measurement conveniently and that the apparatus involved is somewhat clumsy.

The second method is outlined in Fig. 11. The triangular wave may be obtained with a commercial triangular wave generator and the current output is a square wave the rise time of which is dependent on the RC product of the network. By routine circuit analysis the rise time and amplitude of the current step is

$$i_0 = aC(1 - e^{-t/RC}) \tag{22}$$

where a is the slope of one segment of the triangular wave. Thus, if C or R is small enough, a satisfactory current square wave may be generated. The method becomes more interesting in view of (3) which stated that in the feedback circuit the input resistance is further reduced by loop gain. Thus, for example, if we assume that (22) may be applied to the feedback circuit of Fig. 2, with (3) giving the actual input resistance, we may make the following sample calculation. Let us assume that k equals 1000, R_1 is 10^{12} ohms (Fig. 2) and C (Fig. 11) is 1 $\mu\mu$f. Now by (3) the R_{in} of Fig. 11 is 10^9 ohms so that by (22) the time constant of the step function rise is 1 msec; and if a equals 1 volt per second the dc value of the current (equal to ac) is 10^{-12} amperes. This rise time at 10^{-12} amperes would be more than satisfactory.

For an exact analysis refer to Fig. 12. This is the same as Fig. 1 except C_3, the current injecting capacitor, is added. Fig. 13 gives the equivalent circuit which leads to the following equations:

$$e_{\text{in}} = i_1\left(\frac{1}{pC_3} + \frac{1}{pC_1}\right) - i_2\frac{1}{pC_1}$$

$$-e_0 = -i_1\left(\frac{1}{pC_1}\right) + i_2\left(\frac{1}{pC_1} + \frac{R_1}{1 + pC_2R_1}\right). \tag{23}$$

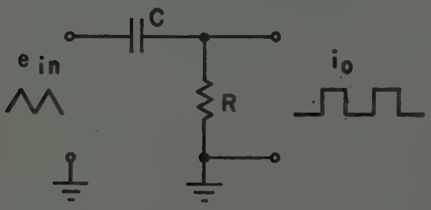


Fig. 11—Triangular wave method of generating a current square wave.

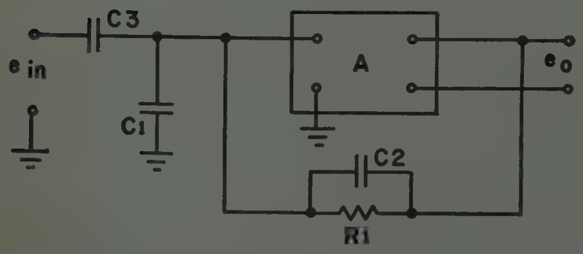


Fig. 12—Micromicroammeter with current injecting capacitor C_3.

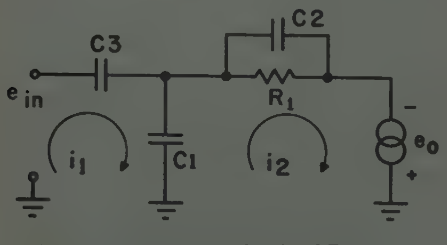


Fig. 13—Equivalent circuit of Fig. 12.

Solution of these equations and stimulation of the network by $e_{\text{in}} = at$, where a is the slope of the ramp function and t is the time, yields, by the Laplace transform method, the result

$$e_0 = \frac{-aR_1C_3k}{k+1} \cdot \left[1 - e\left(\frac{-t}{R_1\left[\frac{C_3}{k+1} + \frac{C_1}{k+1} + C_2\right]}\right)\right] \tag{24}$$

and the time constant is

$$T = R_1\left[\frac{C_3}{k+1} + \frac{C_1}{k+1} + C_2\right]. \tag{25}$$

Therefore, if C_3 the current injecting capacitor is not larger than C_2, even if k were 1, the error would be no more than 50 per cent. If C_1 is much larger than C_2 or C_3, even with $k = 1$, the error in rise time caused by the presence of C_3 will be entirely negligible. With even a moderate loop gain, a fairly large value of C_3 causes no error, and the time constant measured is essentially that of the normal circuit capacities C_1 and C_2. In practice, therefore, to be sure of no error, C_3 should just be large enough to couple into the circuit sufficient step height for convenient measurement (i_{in} equal to aR_1C_3). At very low currents stray circuit capacity will suffice for coupling the signal.

It should be pointed out that the validity of current testing method must be established beyond question before any specification of transient response may be trusted.

APPLICATION

In 1958 and 1959 the U. S. Naval Research Laboratory employed some fifty of these electrometer amplifiers in the upper air research program. A brief summary of flights already accomplished is given in Table I.

In general, the detectors produce a dc current when exposed to radiation in a selected portion of the X-ray or ultraviolet light spectrum. Thus, as the vehicle spins about its axis, the output of the detector varies in a manner determined by the viewing angle of the detector, the spin rate of the vehicle, and the nature of the body producing the radiation. Typical vehicle spin rates have been 1 to 10 rps. In many cases the detector viewing angle has been quite broad allowing the use of circuits with 20 to 50 cps bandwidth. On several occasions, however, narrow collimation has been dictated by the nature of the experiment. One such case was a proportional counter experiment which required an amplifier with a bandwidth of 300 cps and a full scale sensitivity of 10^{-10} amperes.

Circuit performance in flight has, in general, been satisfactory. On two occasions in Aspan flights a noticeable deterioration in frequency response was observed beginning at about 120 seconds of flight and continuing to impact. The apparent cause of this deterioration was

TABLE I

SUMMARY OF ROCKET AND SATELLITE FLIGHTS USING ELECTROMETER CIRCUITS

Vehicle Type	Detectors	Electrometer f.s. Sensitivity	Number of Electrometer Circuits
Vanguard Eta	ion chamber	2×10^{-11} A	1
Explorer VII	ion chamber	2×10^{-11} A	1
Aerobee	ion chamber	2×10^{-11} A	3
	ion chamber	5×10^{-12} A	5
Aspan	ion chambers,	2×10^{-11} A	4
	photomultipliers,	1×10^{-7} A	4
	proportional counters	6×10^{-12} A	19
		1×10^{-10} A	4
		1×10^{-11} A	1
		5×10^{-10} A	1
		5×10^{-11} A	5

transistor changes resulting from excessive temperature rise of the package. In these flights germanium transistors had been used.

Some interference from RF fields has geen encountered in complete flight instrument packages. Usually this type of interference has been eliminated by using ceramic bypass capacitors on each lead entering the electrometer amplifier except for the signal input lead which is shielded.

ACKNOWLEDGMENT

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The Hydrogen Quartector®—A New Phase Detector for Exotic Liquid-Gas Systems*

R. L. BLANCHARD†, SENIOR MEMBER, IRE, AND A. E. SHERBURNE†, MEMBER, IRE

Summary—**The problem of an instrument to discriminate automatically between liquid and gaseous hydrogen has been solved. The principle of operation involves the dependence of acoustic radiation resistance upon density of the medium. The instrument comprises a quartz piezoelectric vibrator immersed in the medium and developing at its terminals a radiation resistance, and an electronic unit external to the medium, which automatically converts this resistance to a signal for measurement or control. The system is suitable for use with most fluids, including hydrogen, oxygen, and kerosene. Hydrogen is of particular interest because low density provides a less distinct interface, particularly at high pressure.**

The radiation resistance of practicable sensor configurations is analyzed for a number of fluids and experimental data are presented for conditions of liquid, gas, and wet in gas. An electronic technique for precise conversion of radiation resistance to a control signal is described. A stable negative resistance is generated, and maintenance of oscillation is the criterion for radiation resistance greater or less than the prescribed value of negative resistance. This negative resistance is chosen to be intermediate in value between the values which obtain for liquid and gas phases. Stable radiation resistance resolution of a few per cent has been obtained easily without encountering any limiting condition.

An example of equipment designed for missile environmental conditions is described.

LIST OF SYMBOLS (MKS UNITS)

Electrical

C = capacitance
E = potential
L = inductance
L_c = intrinsic crystal inductance
L_r = inductance caused by acoustic radiation
Pav = average power, watts
Q = ratio of reactance to resistance
R = net resistance
R_c = intrinsic resistance of mounted crystal
R_n = negative resistance absolute value
R_p = positive resistance
R_r = resistance caused by acoustic radiation
ω = frequency radians per second
X = reactance
Z = impedance.

Piezoelectric

d_{11} = piezoelectric strain coefficient, 2.3×10^{-12} meters/volt
s_{11} = compliance coefficient, 1.27×10^{-11} meters/newton

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l_x = crystal dimension along x axis
l_y = crystal dimension along y axis
l_z = crystal dimension along z axis.

Acoustic

c = velocity of propagation
D = piston diameter
J_1 = Bessel function
K_1 = Bessel function
Pav = average power, joules/second = watts
R_a = acoustic resistance
S = piston area
v = velocity
x = dimensionless parameter
X_a = acoustic reactance
Z_a = acoustic impedance, ratio of force to velocity
ρ = density of the medium
ξ = normalized acoustic resistance
η = normalized acoustic reactance.

I. Introduction

In the control of fuel for liquid-propelled missiles and rockets, need has arisen for a switch to discriminate automatically and reliably between liquid and gas phases. Kerosene, liquid oxygen, and liquid hydrogen flowing in pipes or contained in tanks, and operation on the ground or in flight, typify the environmental conditions.

II. Principle of Operation

The phenomenon on which the system is based is the acoustic radiation resistance of a small piezoelectric vibrator immersed in the medium. This resistance is a function of ρ and c, both of which are different in liquid and gas phases. The functional dependence is determined by the design of the sensor, as will be brought out in Section III, and the sensor can be designed for large discrimination over wide operating conditions. The radiation resistance is a mechanical resistance developed at the "piston" surfaces of the vibrator and is converted by the piezoelectric phenomenon into a corresponding electrical resistance. At the terminals of the piezoelectric crystal, an electrical impedance is observed which is a combination of the equivalent acoustic radiation resistance and reactance, the equivalent electrical resistance of the crystal due to internal losses, the equivalent electrical inductance and capacitance corresponding to the mass and elasticity of the crystal, and the electrical capacitance between the electrodes. All of these elements of the sensor impedance are shown in Fig. 1. Of these, the radiation resistance is a strong function of the properties of the medium in which the sensor is immersed. The remaining part of the system comprises a mechanism for automatically sensing the value of this resistance.

The technique for sensing radiation resistance R_r comprises the generation of the negative admittance shown in Fig. 2, where C_n and R_n represent the absolute magnitudes of a negative capacitance and negative resistance respectively. For the condition $C_n \cong C_e$, the parallel connection of Figs. 1 and 2 is represented by the circuit of Fig. 3. Oscillations are sustained in this circuit in accordance with the criteria:

$$R_p = R_c + R_r$$

$$\text{oscillate } R_n > R_p$$

$$\text{not oscillate } R_n < R_p.$$

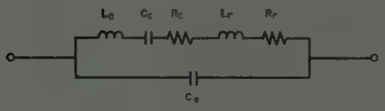

Fig. 1.

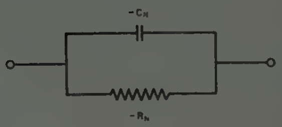

Fig. 2.

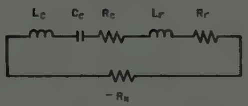

Fig. 3.

Since R_c may be designed to be much smaller than R_r, R_n is a very sensitive criterion for the magnitudes of R_r. If R_n is chosen to be intermediate between the values of R_r for liquid and gas phases of some medium, oscillations are sustained in the gas phase but not in the liquid phase. In the circuit of Fig. 3, the impedance across any element may be described mathematically as a simple pole in the complex plane lying in the left half plane when the sensor is in liquid and in the right half plane when the sensor is in gas.

The presence or absence of oscillation in the circuit is easily converted into a corresponding open or closed position of a relay or transistor switch for indication or control.

Speed of Response

The parameters in Fig. 3 determine the rate of growth or decay of oscillations. For small amplitudes of oscillation R_n is constant, and the relay circuit is designed to operate in this region. At some higher amplitude, R_n diminishes with increasing amplitude, and a stable equilibrium is achieved. In the region of linear operation the oscillations are described as follows:

$$a = a_0 e^{\sigma t} \sin wt = A_{(t)} \sin wt \tag{1}$$

$$\sigma = -R/2L = -w/2Q \tag{2}$$

$$R = R_r + R_c - R_n \tag{3}$$

$$L = L_c + L_r. \tag{4}$$

The time interval for growth or decay of oscillations is described as follows:

$$A_{(t2)}/A_{(t1)} = K = e^{\sigma(t_2-t_1)} = e^{\sigma\Delta t} \qquad (5)$$

$$\Delta t = \frac{1}{\sigma}\log_e K. \qquad (6)$$

Log K and σ are positive for increasing oscillations (gas), and negative for decreasing oscillations (liquid). To determine the switching time, the values are inserted in (5) for $A_{(t2)}$ and $A_{(t1)}$ equal respectively to the relay closure value and the random noise voltage in the absence of oscillations. An oscillogram showing the rapid growth and decay of oscillations for circuit parameters representing the case of hydrogen, liquid and gas, at atmospheric pressure is shown in Fig. 4. A representative time interval for growth of oscillations for this case is somewhat less than 100 msec and for oxygen somewhat less than 10 msec. The corresponding times for decay of oscillation are smaller.

Thus the response time is dependent upon the medium in an interesting way. The value of R in (2) and (3) is the difference between R_n and R_p. If R_n is chosen intermediate between the values of R_p for liquid and gas, then R is one half the difference between the resistance in liquid and the resistance in gas. The smaller the difference between liquid and gas resistances, the longer the time of response for a given sensor design. By making R in (3) very small, the time Δt in (6) can be arbitrarily long. In order to illustrate the growth and decay of oscillations, the following experiment was performed. A sensor with resistance R_p between 1000 and 1100 ohms and with inductance $L \cong 36h$ was connected to negative resistance R_n chosen to have values in sequence $R_{n1} = 1100$ and $R_{n2} = 1000$ ohms.

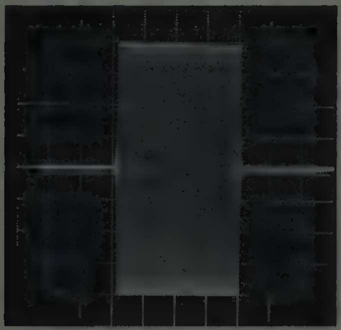

Fig. 4—Speed of response hydrogen switching condition.

Thus

$$|R_{n1} - R_p| + |R_p - R_{n2}| = 100.$$

The circuit Q, therefore, was approximately 500,000. The oscillogram in Fig. 5 where the sweep rate was one vertical line per second shows the corresponding growth and decay of oscillations. Values of K and Δt determined from the oscillogram and (2) and (6) yield the following magnitudes:

$$|R_{n1} - R_p| \cong 39\Omega$$

and

$$|R_p - R_{n2}| \cong 52\Omega.$$

The sum of these (91 ohms) approximates the experimental value, allowing for the approximate value for L.

III. ACOUSTIC RADIATION AND ELECTRICAL RESISTANCE

The vibration of a piezoelectric sensing element radiates sound energy into the surrounding medium. This energy is supplied by the alternating electric field appearing across the crystal. The medium tends to restrain the crystal vibrations, thus requiring a greater voltage to achieve the same extensional amplitude compared to the unloaded crystal. If the crystal extension is maintained constant while the restraining force is varied, then it is clear that the acoustic power radiated is proportional to the force. For a given crystal, the voltage required to produce a given extension is proportional to the force required. Since the electrical power must equal the radiated power, the electrical resistance of the crystal must be proportional to the restraining force.

The reaction of a medium on the vibrating crystal depends on the density of the medium, the velocity of sound in the medium, and the dimensions of the crystal. The loading effect is similar to the acoustic loading on a vibrating piston. The solution for a cylindrical piston in an infinite baffle is well known:

$$Z_a = \rho c S[\xi + j\eta] = R_a + jX_a \qquad (7)$$

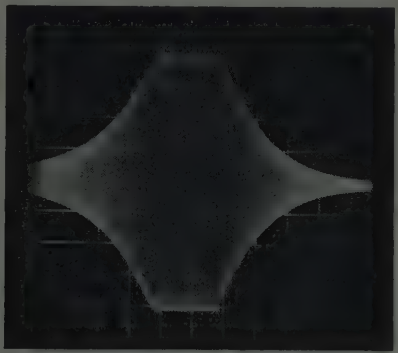

Fig. 5—Growth and decay of oscillations.

where

$$\xi = 1 - \frac{2J_1(x)}{x} = \frac{x^2}{2\cdot 4} - \frac{x^4}{2\cdot 4^2\cdot 6} + \frac{x^6}{2\cdot 4^2\cdot 6^2\cdot 8} - \cdots \qquad (8)$$

$$\eta = \frac{2K_1(x)}{x^2} = \frac{4}{\pi}\left(\frac{x}{3} - \frac{x^3}{3^2\cdot 5} + \frac{x^5}{3^2\cdot 5^2\cdot 7} - \cdots\right). \qquad (9)$$

In this paper we are primarily interested in R_a, since this appears as a resistance in the electrical equivalent circuit. Although X_a appears in the equivalent circuit as a reactance, it serves only to cause a slight frequency shift in the oscillator.

In (7), ρc is the characteristic impedance of the medium per unit area, S the area, and the term in brackets characterizes the size and shape of the piston in terms of a wavelength in the medium. For a circular piston, $x = wD/c$. For the rectangular piston used in the design described in this paper, we have used the dimension 1.7 l_x in place of D where l_x is the smaller of the two piston dimensions, giving:

$$x = \frac{1.7 w l_x}{c} \cdot \tag{10}$$

This estimated value for x has been used in lieu of performing a pressure integration over the piston face, and for the requirements of the application has been quite adequate.

It is interesting to observe the effect of sensor dimensions on the functional dependence of R_a upon ρ and c. For x large, *i.e.*, for a piston in which the smallest dimension is larger than a wavelength in the medium, R_a is roughly proportional to ρc. For x small, R_a is roughly proportional to ρ/c as the following development shows: For $x<1$, it is convenient to use the power series form for Z_a:

$$R_a = \rho c S \left[\frac{\left(\frac{2\pi f D}{c}\right)^2}{2\cdot 4} - \frac{\left(\frac{2\pi f D}{c}\right)^4}{2\cdot 4^2\cdot 6} + \cdots \right]$$

$$\cong \frac{(2\pi)^2 D^2 f^2}{8} \frac{\rho}{c} S \left(\frac{2\pi f D}{c} < 1 \right) \tag{11}$$

and

$$X_a = \frac{4}{\pi} \rho c S \left[\frac{\left(\frac{2\pi f D}{c}\right)}{3} - \frac{\left(\frac{2\pi f D}{c}\right)^3}{3^2\cdot 5} + \cdots \right]$$

$$\cong \frac{8Df\rho}{3} S \qquad \left(\frac{2\pi f D}{c} < 1 \right) \tag{12}$$

where f is the frequency of the radiated energy.

The preceding equations describe a mechanical impedance at the piston surface. The corresponding electrical impedance which is developed in the sensor equivalent circuit (Fig. 1) can be obtained by the following argument.

The acoustic power radiated is:

$$\text{Pav} = \tfrac{1}{2} R_a V_0^2 \tag{13}$$

where the piston velocity is:

$$v = V_0 \cos wt. \tag{14}$$

The electrical power supplied to the crystal is:

$$\text{Pav} = \frac{E^2}{2R_r} \tag{15}$$

where R_r is the electrical resistance corresponding to the acoustic radiation resistance R_a and E is the peak applied voltage. Setting the two expressions equal and solving for R_r:

$$R_r = \frac{E^2}{V_0^2 R_a} \cdot \tag{16}$$

The velocity of the end of a crystal driven by a sinusoidal voltage $E \cos wt$ is:

$$v = \frac{d_{11} S}{s_{11} l_x R_a} E \cos wt. \tag{17}$$

Therefore

$$V_0 = \frac{d_{11} S E}{s_{11} l_x R_a} \cdot \tag{18}$$

Solving for E/V_0 and substitutiug in the expression for R_r yields

$$R_r = \frac{s_{11}^2 l_x^2}{d_{11}^2 S^2} R_a \tag{19}$$

$$R_r = \frac{s_{11}^2}{d_{11}^2} \frac{l_x}{l_z} \rho c \left(1 - \frac{2J_1(x)}{x} \right). \tag{20}$$

Eqs. (10) and (20) have been used in subsequent computations.

For $x<1$, the approximate expression for R_r corresponding to (20) becomes:

$$R_r \cong (2\pi)^2 \frac{s_{11}^2}{d_{11}^2} \cdot \frac{(1.7)^2 \cdot {}^2 l_x^3}{8 l_z} \frac{\rho}{c} \cdot \tag{21}$$

The resistance for both crystal faces is one half this value

IV. Impedances of the Sensor in Liquids and Gases

The impedance of the sensor has been computed and measured for a variety of gases and liquids. Measurements of the impedance magnitude in liquids was performed with a transmission circuit which uses the sensor in a voltage-dividing network. The sensor impedance magnitude is thus compared to a known resistance. Resistance measurements have been made by an alternate technique for the case of liquid hydrogen and for several gases. In this experiment, the values for R_n for which oscillations grow and decay are measured. These data are in agreement with the transmission circuit data.

Table I gives computed radiation resistance values in electrical ohms for several liquids and gases. Eq. (20) was used for all calculations, with D taken as 0.070 inch. Data for density and velocity of sound were taken from several sources for which references are given at the end of Table I. These data are in conformance with the experimental data now to be described within the accuracy to which calculations and measurements were attempted.

The impedance magnitude of the sensor in liquid nitrogen experimentally is shown in Fig. 6. The curve is

TABLE I

Calculated Radiation Resistances of Gases and Liquids

Gas	Temperature, °K	Pressure, atmospheres	Radiation Resistance, ohms
Argon[1]			
	300	1	800
	300	10	8,100
	300	40	33,000
	300	70	59,000
	300	100	84,000
Helium[2]			
	255	1	50
	255	100	4,900
Nitrogen[1]			
	100	1	1,300
	200	1	830
	300	1	550
	300	10	5,600
	300	100	53,000
Oxygen[1]			
	120	1	1,300
	300	1	650
	300	10	7,000
	300	100	59,000
Liquids			
Hydrogen[3]	20	1	15,000
Nitrogen[3]	77	1	220,000
Oxygen[3]	90	1	310,000
JP-4[2]	298	1	160,000
UDMH[4]	298	1	160,000

plotted starting with a dry sensor and slowly lowering it into the liquid. The impedance jumps suddenly when the crystal is still about $\frac{1}{16}$ inch above the liquid because surface tension causes some liquid to flow up the support wires and wet the crystal. The impedance rises rapidly until the crystal is submerged about $\frac{1}{10}$ inch below the surface. A slight oscillation in the impedance with increasing depth is caused by reflections from the surface. This effect disappears as the sensor depth is increased slightly. When the sensor is removed, some of the liquid adheres and the sensor resistance is considerably higher than that of the dry sensor. Evaporation of the liquid finally returns the impedance to its original value. The curve of Fig. 6 is typical of curves obtained for all liquids except that the wet and submerged values, of course, are different for different liquids. Fig. 7 shows the range or impedances measured for several liquids. The wet impedances as well as the submerged impedances are shown. The bars marked S_1 and S_2 represent the performance of two switches, one designed for liquid oxygen and the other for liquid hydrogen. The extremities of each bar show the range of impedance values for which switching is assured for interchangeable units over military environmental conditions. Notice

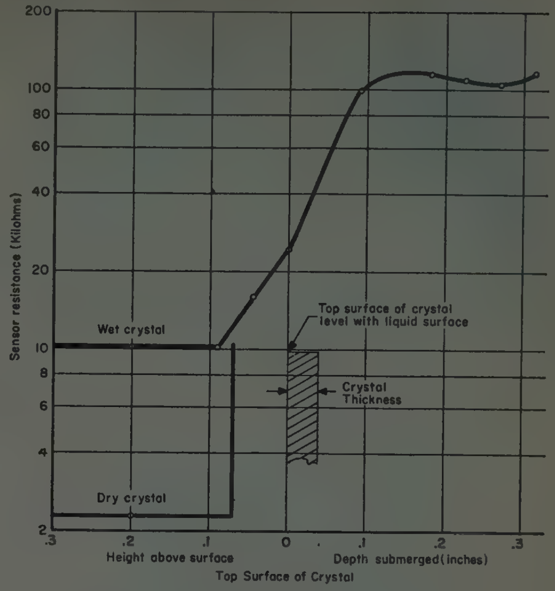

Fig. 6—Sensor impedance magnitude in liquid N_2 atmospheric pressure.

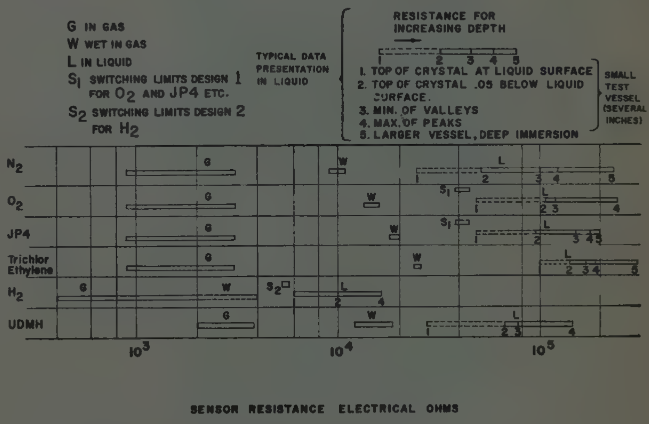

Fig. 7—Sensor impedance magnitude in liquids.

that, in both cases, the switching limits are well removed from the extreme values measured for 0.05-inch submersion on one side and for a wet crystal in gas on the other.

Experimental data for the impedance of the sensor in gases are given in Fig. 8. The rising characteristic of the curves for helium and nitrogen is believed to be caused by a reflection of acoustic waves from the glass header of the sensor and the variation in the velocity of sound with pressure. Note that the resistance which is plotted is the sum R_c+R_r. For the sensor used in these measurements, R_c was approximately 800 ohms. These data were obtained from transmission measurements and verified by the measurement of the negative resistance required to cause oscillations.

V. Electronic Discrimination of Impedance

Automatic classification of sensor impedances as greater or less than a prescribed value is accomplished by two functions:

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[3] R. B. Scott, "Cryogenic Engineering," D. Van Nostrand Co., Inc., Princeton, N. J.; 1959.

[4] Anonymous, "Storage and Handling of Dimazine (unsym-Dimethylhydrazine)," prepared by Res. and Dev. Dept. of Westvaco Chlor-Alkali Div., Food Machinery and Chemical Corp., New York, N. Y., 3rd ed.

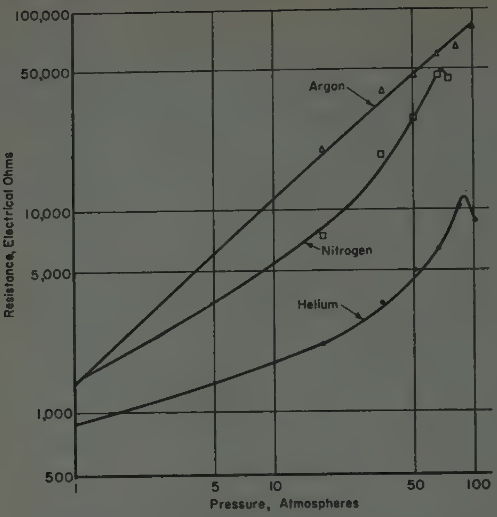

Fig. 8—Sensor resistance vs pressure for several gases at approximately 70°F.

1) generation of a negative admittance of prescribed value; and
2) detection of oscillation and operation of a control circuit (relay or transistor switch).

Fig. 9 illustrates the generation of negative admittance. It comprises an amplifier with gain A over a bandwidth large compared to the bandwidth of the circuit in Fig. 3. This network generates the negative admittance indicated in Fig. 2 in accordance with (22)–(24).

$$\frac{1}{Y_n} = -\frac{Z_f}{1+\left(\dfrac{Z_f+2R_1}{AR_1}\right)}\left[1-\frac{1}{A}\left(2+\frac{3R_1}{Z_f}\right)\right] \qquad (22)$$

$$C_n \cong C_f \qquad (23)$$

$$R_n \cong R_f\left[\frac{1}{1+\dfrac{R_f}{AR_1}}\right]. \qquad (24)$$

For a large and stable value of AR_1, stable values of R_n are obtained over a wide range of values by choice of R_f. An illustration of this is given in Fig. 10 where (24) for R_n vs R_f is plotted. Also shown in the lower part of the illustration is the result of an experiment where for a given value of R_f, the values of a positive resistance R_p, for which the circuit does or does not oscillate and operate a control circuit, are plotted. This is indicated for eight values of R_f, and the differences between these values and the computed value of R_n are shown. In the experimental equipment in which these data were contained, R_p was known only to an accuracy of about one per cent, and the smallest step which could be made in R_p was 10 ohms. Thus, the placement of the bars on the page include about a one per cent uncertainty due to experimental equipment.

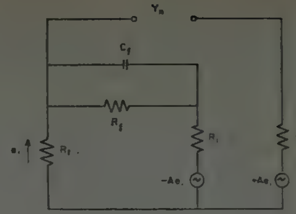

Fig. 9.

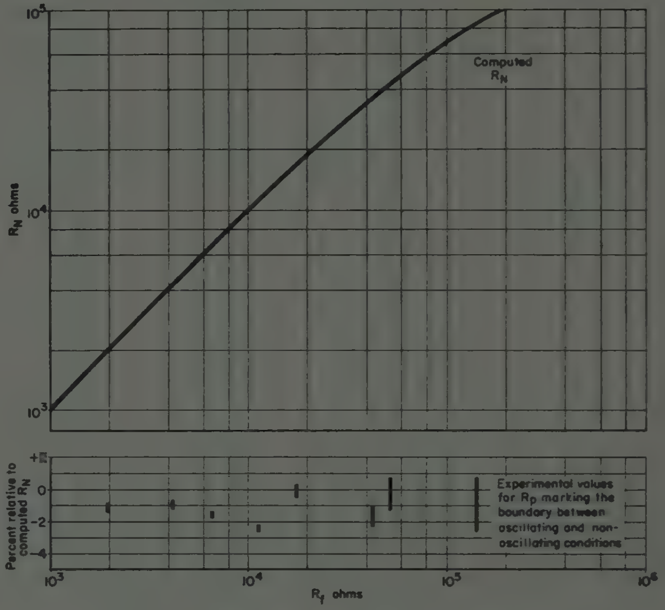

Fig. 10—Electronic unit resistance resolution.

Although electronically generated, a high degree of stability and predictability is achieved in the negative resistance R_n. In (25), the dependence of R_n upon R_f and upon $R_a = AR_1$ is shown.

$$\frac{\delta R_n}{R_n} = \frac{1}{1+\dfrac{R_f}{R_a}} \cdot \frac{\delta R_f}{R_f} + \frac{R_n}{R_a}\left[\left(\frac{\delta R_a}{R_a}\right)_i + \left(\frac{\delta R_a}{R_a}\right)_v + \left(\frac{\delta R_a}{R_a}\right)_t\right]. \qquad (25)$$

The subscripts i, v, and t refer to interchangeability, voltage, and temperature respectively. From both design and experimental data, the components of (25) have been derived and the result is plotted in Fig. 11. This degree of stability is representative of a practicable design and is entirely adequate for the switching application described in this paper. For more demanding applications, greater stability is readily achievable. The environmental conditions which apply to the data in Fig. 11 include interchangeability of units, voltage variation between 25 and 31 volts and temperature variations from −50°C to +75°C.

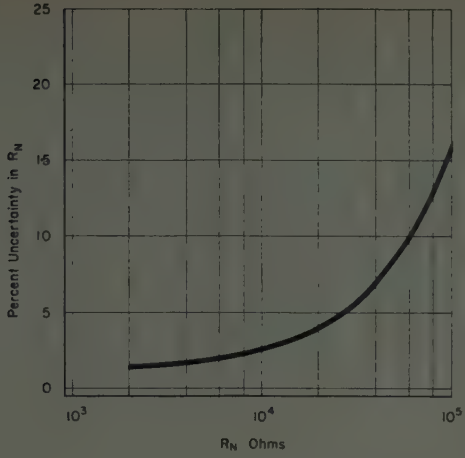

Fig. 11—Per cent uncertainty in R_n.

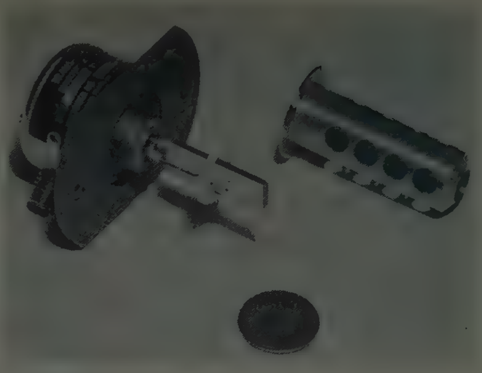

Fig. 12—"In tank" sensor, cage removed.

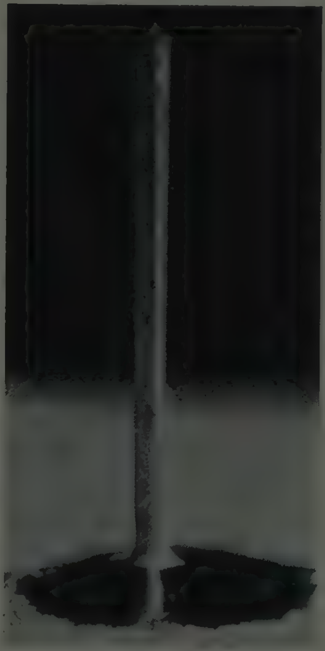

Fig. 13—Support lead attachment to crystal surface about 25×.

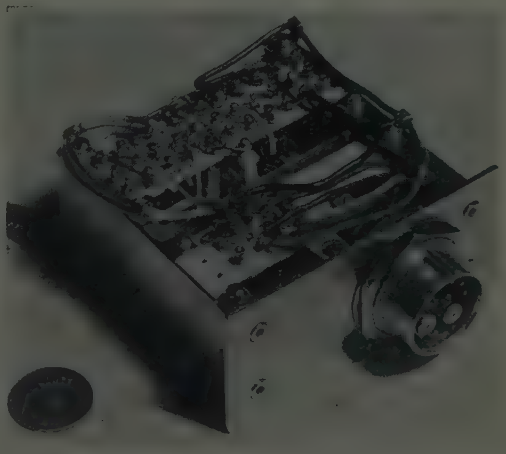

Fig. 14—Model of electronic unit without case.

VI. Sensor Design

The basic element of the sensor is a piece of natural crystalline quartz, x cut (+5°) which vibrates in a direction approximately parallel to the longest axis. It is approximately one inch long and radiates acoustic energy from both end surfaces.

The quartz plate is supported by four leads, two pairs on each major surface. The leads are headed beryllium copper wires, vacuum heat treated, gold plated and bonded to the quartz surface by epoxy resin. The leads are mounted off the nodal line which runs between the leads and approximately normal to the long axis of the crystal. This location of the leads provides an excellent mechanical mounting by providing a supporting couple about the axis of greatest moment of inertia. The lead length is chosen to provide a broad resonance around the vibrating frequency (~100 kc), and a correspondingly low mechanical impedance at the crystal surface. The mounting permits Q's of 30,000 or so. A general view of one version of this design is shown in Fig. 12, and Fig. 13 is a magnified view of the support lead attachment to the crystal surface. The unit shown in Fig. 12 has electrodes of silver. Another version suitable for corrosive fluids is geometrically similar but platinum has been substituted for silver, metal bonding for epoxy resin, etc. These designs are very rugged and will withstand and operate under sinusoidal accelerations of 25 g at frequencies up to 2000 cycles.

VII. Electronic Unit Design

A model of one design of the electronic unit is shown in Fig. 14. The amplifier portion comprises three transistor stages having a total gain of 72 db. By choice of value of resistor R_f, it is capable of generating negative resistances between about 1500 and 200,000 ohms. Both transistor and relay-type control circuits have been used with this type of amplifier. The model shown includes a control circuit comprising a single transistor dc amplifier and a relay. The production units of the same design are interchangeable and may be used without adjustment with arbitrary cable lengths between electronic unit and sensor from 0 to 250 feet.

VIII. Conclusion

A thoroughly reliable and highly discriminating technique for distinguishing between liquid and gaseous phases of fluids has been described. Conditions of splashing and a wet sensor in gas are easily distinguished from the in-liquid conditions. The components and the phenomenon are compatible with a wide variety of fluids including cryogenic propellants. The technique used includes the stable generation of negative resistance and provides a technique for the measurement of radiation resistance which may have additional applications.

IX. Acknowledgment

The authors acknowledge the contributions of V. C. Westcott, who originally suggested the idea of utilizing the damping of a crystal when immersed in a liquid, and of R. Arsenault, J. Meyer, and J. Middleton, who have contributed to the design, construction, and testing of the equipments described here, and of Prof. R. B. Belser of Georgia Institute of Technology for the photograph in Fig. 13.

An Instrument for Measuring Liquid Level and Slosh in the Tanks of a Liquid-Propellant Rocket*

L. B. WILNER†, W. L. MORRISON†, MEMBER, IRE AND A. E. BROWN†, MEMBER, IRE

***Summary*—A measuring system for monitoring dynamic level variations in red fuming nitric acid is discussed. The system employs a continuous variable-capacitance sensor with a tuned-circuit capacitance detector. The transistorized, temperature-stabilized capacitance detector is discussed in detail. Performance data are presented.**

Background

As part of the diagnostic instrumentation for use in the development of liquid fuel rockets, an instrument was required which could monitor the level and surface attitude in the rocket propellant tanks. The original requirement written for this instrument was for monitoring only an oxidizer, red fuming nitric acid. It is only this instrument, called the Oxidizer Slosh Monitor, which will be discussed here.

Before proceeding further, it is worthwhile to consider the properties of the material in which the monitor was to function. Red fuming nitric acid is a liquid, slightly more dense and slightly more viscous than water. Extremely corrosive, it reacts with most organics and dissolves most metals, though at varying rates. Electrically, the acid is about as conductive as a saturated salt solution, but the conductivity of acid acceptable as propellant varies over a considerable range.

The device for monitoring this material was to provide a continuous measure of level along three or more lines up and down the curved wall of a tank. Thus, at any time, the mean level and the attitude of the liquid surface relative to the rocket axis might be determined, assuming the liquid surface remained a plane. Because sloshing of the propellant might be of great importance in diagnosing a rocket failure, the monitor had to be capable of following level oscillations up to 3 cps.

One of the first techniques tried in developing a monitor was the use of a high-resistance wire, shorted to the tank wall by the liquid at the liquid surface and below. As so often happens with such simple techniques there were unforeseen complications. Specifically, the action of the acid on the wire made its resistance change and interfered with the electrical contact between the wire and the liquid.

To avoid the effects of acid reacting with the metal wire, an insulating material was interposed between the wire and the acid, changing the sensor to a capacitive element similar to those used in wave gages.[1–3] Various insulating materials were tried, and it was found that of the materials tried, only fused Kel-F insulation retained its electrical properties when immersed in nitric acid.

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* Original manuscript received by the IRE, December 2, 1959. Work performed under Air Force Contract AF04(647)-347.

† Missiles and Space Div., Lockheed Aircraft Corp., Palo Alto, Calif.

[1] W. S. Campbell, "An Electronic Wave-Height Measuring Apparatus," David Taylor Model Basin, Washington, D. C., Rept. No. 859; October, 1953.

[2] G. R. Barnard, "An Evaluation of Wave-Height Measuring Equipments," Instrumentation Div., David Taylor Model Basin, Washington, D. C., Tech. Note No. 9; September, 1958.

[3] C. G. Whittenbury, E. A. Huber, and G. S. Newell, "Instrument for measuring water waves," *Rev. Sci. Instr.*, vol. 30, pp. 674–677; August, 1959.

THE CAPACITIVE SENSOR

The sensing element which was used for the Oxidizer Slosh Monitor was a length of wire, insulated with Kel-F, strung in a vertical plane on Teflon stand-offs along the curved wall of the oxidizer (acid) tank. When immersed in the acid, this wire formed a capacitor of which the insulation was the dielectric, the wire was one "plate," and the acid at the outside surface of the insulation was the other "plate." Above the surface of the acid, the wire and the tank wall formed a capacitor of much lower value. Thus, in effect, the sensor was a cylindrical capacitor, the length of which was determined by the oxidizer level.

Typical capacitance values for this sensor were 0.5 $\mu\mu$f/inch between dry wire and the tank wall, and 3.5 $\mu\mu$f/inch for the immersed wire. In the frequency range used (below 5 mc), the impedance of this capacitance was such that the resistance of the acid was comparatively small, and the resistance of the wire was negligible.

FLOWBACK

When the liquid level was oscillating (*i.e.*, under sloshing conditions), the liquid in the immediate vicinity of the sensor wire did not follow perfectly the mean liquid level. Instead, a film of liquid clung to the sensor, causing a phenomenon known as flowback.[1,2] Fig. 1 illustrates graphically the flowback film and its effect on the apparent capacitance of the sensor element. As indicated by the figure, some of the sensor wire actually above the level of the liquid remained electrically connected to the liquid by the flowback film. For a sinusoidal variation of liquid level, the effect of flowback on sensor capacitance, hence on instrument output, was to decrease peak-to-peak amplitude, shift the phase, and distort the waveform by flattening one peak, sharpening the other.

The most serious effect of flowback on the Oxidizer Slosh Monitor was the attenuation of slosh amplitude. For slosh amplitude of the order of a few inches, the attenuation from flowback exceeded 60 per cent at a frequency of only 2 cps in some early models. The attenuation from flowback was aggravated when the mean liquid level crossed a sensor wire support. The wire supports would catch a quantity of liquid at the peak of a slosh and drip it down the sensor wire throughout the rest of the slosh cycle, essentially obscuring the variation of the main liquid surface.

MEANS OF MINIMIZING FLOWBACK

The effects of flowback on slosh frequency response were found to be dependent on at least two variables, the sensor wire diameter and the carrier frequency at which the sensor capacitance was measured.

It was found that as sensor wire diameter was decreased, the slosh frequency response was improved. This effect resulted from the utilization of surface tension to dissipate the flowback film. Unfortunately, the slosh monitor application did not permit the use of a very fine wire. The wire had to be installed in hard to reach places and fed out through a large lid. The mortality on AWG No. 36 wires, which were tried first, was discouragingly high. An AWG No. 30 wire has been used in the more recent applications, although the diameter is somewhat large to make good use of surface tension effects.

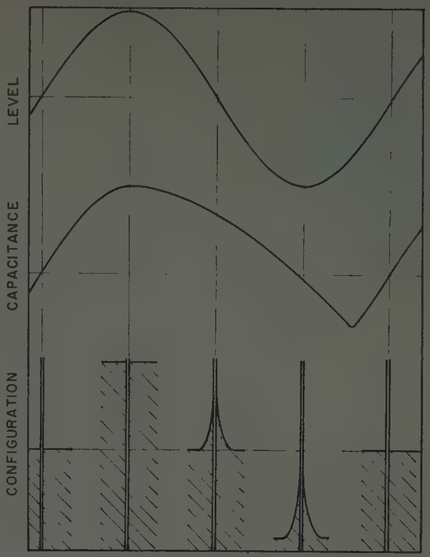

Fig. 1—Flowback and its effect on the output of a variable-capacitance level sensor.

By varying the carrier frequency at which the sensor capacitance was measured, it was possible to differentiate somewhat between wire immersed in the liquid and wire which was only wetted by a flowback film. An element of sensor wire wetted by a flowback film differed electrically from a completely immersed element by having in series with it the resistance of the flowback film. By increasing the carrier frequency, the capacitive impedance of that element could be decreased so that it might be effectively excluded from the circuit by the series resistance of the flowback film.

This principle of differentiating flowback and main level was applied to the Oxidizer Slosh Monitor by increasing the carrier frequency to 3 mc. The carrier frequency was more or less limited at this value because the required length of sensor, acting as a transmission line, became an appreciable fraction of a wavelength at 3 mc.

The use of the 3-mc carrier gave the Oxidizer Slosh Monitor an acceptable slosh frequency response despite the use of relatively large wire (see "Performance" section). However, no means was found of eliminating the degradation of slosh response caused by wire supports in the slosh.

THE ELECTRONICS

Detection of the capacitive changes of the sensor wire is somewhat complicated by the necessity of making measurements at 3 mc. Wide variances in sensor capacitance, the balancing problems and high power levels associated with the majority of capacitive measuring techniques, led to a selection of a slope detection system. This system provided a means of determining, with reasonable linearity, the variation in capacitance with a minimum power consumption. Fig. 2 shows the resultant system in block form.

A crystal-controlled oscillator was selected to afford maximum frequency stability and minimum temperature dependence. The frequency of the oscillator was held constant by the crystal and thereby one variable was eliminated from the system. As a means of eliminating the oscillator amplitude as a variable, a clamp circuit was provided which limited the peak-to-peak amplitude to a fixed value over the environmental range of the equipment. In this manner the input to the slope detector was held constant in frequency and amplitude.

Amplitude dependence on sensor capacitance was developed by use of a tuned circuit which acted as a variable impedance in a voltage divider circuit. The resonant circuit consisted of a transformer which was resonated slightly above the crystal frequency when the sensor capacitance was maximum (the full-tank condition). As the sensor capacitance decreased, the resonant frequency of the transformer increased and the signal amplitude at the resonant circuit decreased. The linearity and capacitance range of the resonant circuit was controlled by the turns ratio of the transformer and the series-parallel capacitor network used in conjunction with the sensor.

Capacitance data in the form of variable amplitude of the 3-mc carrier frequency are converted to a usable form by peak-to-peak detection in a thermistor temperature-compensated detector. The resulting slowly varying dc voltage, dependent on sensor capacitance, was made suitable for injection into a conventional telemetry system by use of an emitter-follower output for impedance matching. Since the output is dependent upon the resonant circuit characteristic curve, it is not linear over the entire range, and therefore a calibration curve was required to convert from dc output voltage to tank liquid level. A circuit diagram of the completed system is shown in Fig. 3.

Performance

The Oxidizer Slosh Monitors which have been built and tested provided acceptable, but not ideal, performance. The response to static level was not linear, but was an "S" curve for which deviation from linearity was less than 5 per cent over the range used. (See Fig. 4.) Typical repeatability of indicated level information over a temperature range of +10°C to +50°C was 1.5 per cent of full scale. The ratio of full-tank output to empty-tank output was about 10:1, which did not approach the 30:1 ratio which the telemetry could accept, but was enough to provide clear definition of level throughout the range. Because of the deviation of the output/level curve from linearity, the sensitivity of the monitor to slosh was different at different liquid levels.

The effects of flowback were not completely eliminated from the system. The output for a 3-inch slosh amplitude declined by about 25 per cent at 3 cps, more at higher frequencies. (See Fig. 5.)

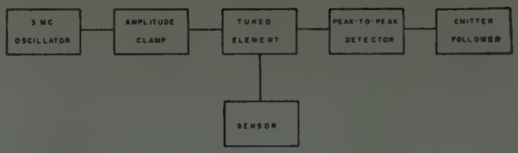

Fig. 2—Block diagram of the Oxidizer Slosh Monitor.

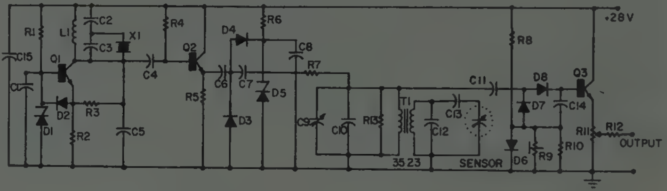

Fig. 3—Circuit schematic of the Oxidizer Slosh Monitor.

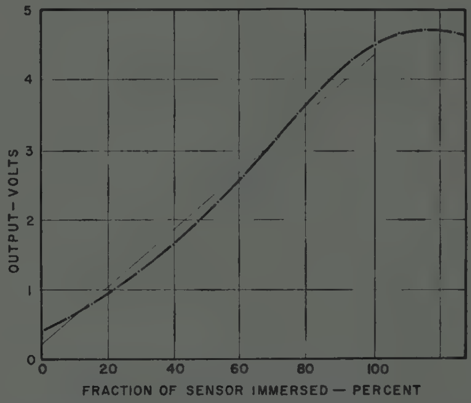

Fig. 4—Output vs sensor immersion for the Oxidizer Slosh Monitor.

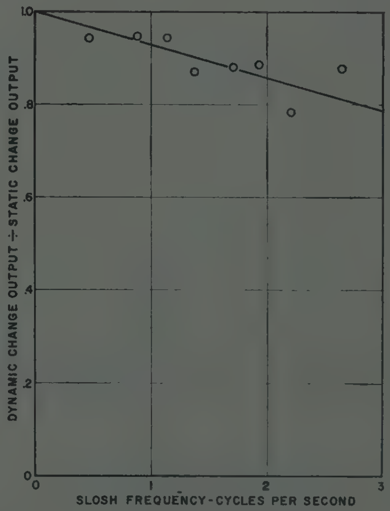

Fig. 5—Slosh frequency response of the monitor in red fuming nitric acid.

Acknowledgment

The authors wish to acknowledge the guidance of Dr. T. A. Perls in developing the concept of the slosh monitor, and the craftsmanship of J. L. Hatchel and H. L. Koehl in assembling the prototype models.

Correspondence

Noise Performance Theory of Esaki (Tunnel) Diode Amplifiers*

In a recent letter, K. K. N. Chang[1] described an amplifier using the Esaki Diode. He presented expressions for the gain g_p and noise figure F of a two-port circuit with separate generator and load impedances. These expressions are

$$g_p = \frac{4G_gG_L}{(G_L + G_g + G_1 - G)^2} \qquad (1)$$

and

$$F = 1 + \frac{T}{T_0}\left(\frac{G_1}{G_g} + \frac{G_L}{G_g} + \frac{G_e}{G_g}\right). \qquad (2)$$

In the above, T is the circuit temperature; T_0 is the reference temperature; G is the diode negative conductance; G_g is the source generator conductance; G_1 is a parasitic circuit-loss conductance; G_L is the load conductance; and G_e is an artificial electronic conductance $eI_0/2kT$ in which I_0 is the direct current of the diode, k is Boltzmann's constant and e is the electronic charge. (Chang also included a transformer turns-ratio which appeared in his circuit. This has been eliminated as unnecessary to the theory. If transformers are to be used, they can be considered to modify the values of G_L and G_g, but not G_1 as this parasitic resistance tends to be associated with the diode itself.) The above expression for noise figure is based upon the assumption that the diode is characterized by pure shot noise and that the resistances are sources of thermal noise only.

Direct application of (2) indicates that a very low noise figure could be obtained by raising G_g to a large value. This, however, cannot be done without considering the effect upon gain, which depends upon the negative conductance G, a parameter not included in (2). When this effect is taken into account it will be seen that useful low-noise amplification depends upon obtaining a low value for the ratio $eI_0/2kTG$ and for the ratio G_1/G. In Chang's discussion in his letter and in his paper at WESCON, 1959, he also implies that $eI_0/2kTG$ is the most important parameter but does not show the manner in which this expression appears.

We shall present here an analysis of the one-port form of amplifier which is, in general, superior to the two-port variety for low noise. Fig. 1 shows the circulator method of interconnecting one-port amplifiers. Fig. 2 shows a simplified equivalent circuit for a single amplifier and its connecting transmission line. If a transformer is to be used it should be at the input of this circuit, which has the effect of allowing G_g to assume any desired value as far as this analysis is concerned.

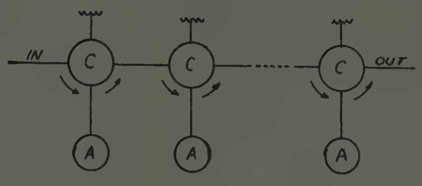

Fig. 1—Tandem arrangement of one-port amplifier A and circulators C.

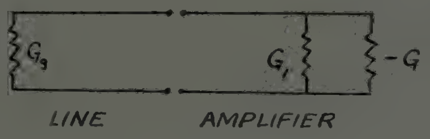

Fig. 2—Simplified equivalent circuit of one-port amplifiers.

The gain of the one-port amplifier is given by the square of the magnitude of the reflection coefficient on the input line. This is

$$g_p = \left[\frac{G_g - G_1 + G}{G_g + G_1 - G}\right]^2, \qquad (3)$$

and the noise figure (for the above assumptions) is

$$(F)_{\text{one port}} = 1 + \frac{4T}{T_0}G_gG\,\frac{\left(\frac{eI_0}{2kTG} + \frac{G_1}{G}\right)}{(G_g - G_1 + G)^2}. \qquad (4)$$

$$(F)_{n\text{ stages two port}} = 1 + \frac{T}{T_0}\left\{\frac{1 - \frac{1}{g_p{}^n}}{1 - \frac{G_1}{G}}\right\}\left(\frac{eI_0}{2kTG} + \frac{G_1}{G} + \frac{G_L}{G}\right). \qquad (10)$$

As in (2) we can make $(F-1)$ approach zero by allowing G_g to become arbitrarily large. This, however, is no help because the gain is reduced in the process and then several stages may be needed, each of which contributes to the over-all noise figure of the system. Consider (4) when the gain is high. Then G_g approaches $(G-G_1)$ and the expression becomes

$$(F)_{\text{high gain one port}} = 1 + \frac{T}{T_0}\left\{\frac{1}{1 - \frac{G_1}{G}}\right\}\left(\frac{eI_0}{2kTG} + \frac{G_1}{G}\right). \qquad (5)$$

If the gain is not high we must consider two or more stages in tandem. In this case, the over-all noise figure for n stages can be written

$$(F)_{n\text{ stages}} = 1 + M\left(1 - \frac{1}{g_p{}^n}\right), \qquad (6)$$

where M is the "noise measure" of one stage defined by Haus and Adler[2] as

$$M = \frac{F - 1}{1 - \frac{1}{g_p}}. \qquad (7)$$

Thus the net noise figure for n stages becomes

$$(F)_{n\text{ stages one port}} = 1 + \frac{T}{T_0}\left\{\frac{1 - \frac{1}{g_p{}^n}}{1 - \frac{G_1}{G}}\right\}\left(\frac{eI_0}{2kTG} + \frac{G_1}{G}\right). \qquad (8)$$

In this expression G_g does not appear except indirectly in that it affects the gain of a single stage and the number of amplifiers required for a given system gain. When the total gain of n stages is high, it is seen that the over-all noise figure for many low-gain stages is the same as for a single high-gain stage.

We can devise a similar expression for a tandem arrangement of two-port amplifiers using (2). In order to minimize M, we have set

$$G_g = G_L + G - G_1. \qquad (9)$$

We have also assumed that the stages are separated by isolators at a temperature T. The resulting over-all noise figure expression becomes (10).

If we substitute (9) into Chang's gain expression (1), we obtain

$$g_p = 1 + \frac{G}{G_L} - \frac{G_1}{G_L}. \qquad (11)$$

If we wish to obtain a low over-all noise figure, (10) indicates that G_L/G must be small, and from (11) this implies that the gain of each stage must be high. Thus, the two-port amplifier can give good noise performance only if the gain is high and the load conductance G_L small. As G_L appears in the numerator of the gain expression (1), this requires the denominator to be small indeed and we must work close to the point of instability. This will also reduce the bandwidth. With two-port amplifiers it is possible, theoretically, to approach the noise performance of one-port amplifiers but only by sacrificing stability and bandwidth.

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* Received by the IRE, September 17, 1959.

[1] K. K. N. Chang, "Low-noise tunnel-diode amplifier," Proc. IRE, vol. 47, pp. 1268–1269; July, 1959.

[2] Haus and Adler, "Circuit Theory of Linear Noisy Networks," John Wiley and Sons, Inc., New York, N. Y.; 1959.

Firing Angle in Series-Connected Saturable Reactor*

H. F. Storm[1] has discussed the series-connected saturable reactor with free even harmonics (see Fig. 1). He states that in the overexcited case, the firing angle α remains constant for a range of values of the control voltage E_c. However, he does not prove this. In a letter to the author, Storm stated he is not aware of any proof of this being published. It is the purpose of this letter to derive the proof.

Consider Fig. 2. During the interval $\omega t < 2\pi$, the control voltage $E_c = 0$, and so the flux curves of ϕ_A and $-\phi_B$ coincide. At $\omega t = 2\pi$, a small control voltage is introduced. Both cores will tend to saturate, but core A will tend to saturate more quickly now than before E_c was applied, and core B more slowly. Thus, core A will saturate at an angle α while core B is as yet unsaturated.

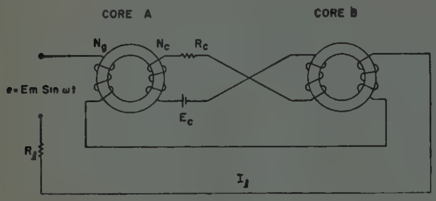

Fig. 1.

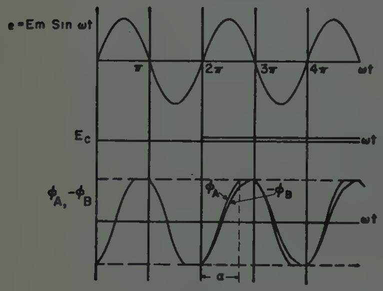

Fig. 2.

The angle α can be determined from (1)

$$2\phi_s = \frac{1}{\omega}\int_0^{\alpha}\left(\frac{E_m \sin \omega t}{2N_g} + \frac{E_c}{2N_c}\right) d(\omega t) \qquad (1)$$

where ϕ_s is the saturation flux of the cores.

Core B will continue toward saturation after core A has become saturated. Depending on the value of E_c, it may or may not saturate during the course of the half cycle (180°). Let us determine the conditions under which core B will saturate at the very end of the half cycle. During the interval that core A is as yet unsaturated, $2\pi < \omega t < \alpha$, the flux in core B will change by an amount given by

$$\Delta\phi_1 = \frac{1}{\omega}\int_0^{\alpha}\left(\frac{E_m \sin \omega t}{2N_g} - \frac{E_c}{2N_c}\right) d(\omega t). \qquad (2)$$

When core A is saturated, the voltage across its primary is zero. Core B acts as a transformer with a secondary resistance R_c. The voltage across the primary of core B is therefore given by

$$E_B = \frac{R_c\left(\frac{N_g}{N_c}\right)^2}{R_l + R_c\left(\frac{N_g}{N_c}\right)^2} E_m \sin \omega t$$

$$\approx \frac{R_c}{R_l}\left(\frac{N_g}{N_c}\right)^2 E_m \sin \omega t \qquad (3)$$

The above approximation is based on the assumption that

$$R_c\left(\frac{N_g}{N_c}\right)^2 \ll R_l. \qquad (4)$$

During the interval that core A is saturated, the change in the flux of core B is given by

$$\Delta\phi_2 = \frac{1}{\omega}\int_{\alpha}^{\pi}\left[\frac{E_m}{N_g}\frac{R_c}{R_l}\left(\frac{N_g}{N_c}\right)^2 \sin \omega t - \frac{E_c}{N_c}\right] d(\omega t). \qquad (5)$$

When E_c is adjusted so that core B saturates at the very end of the half cycle, we have

$$\Delta\phi_1 + \Delta\phi_2 = 2\phi_s. \qquad (6)$$

Integrating and then combining the above equations, using the approximation of (4) again, and defining the voltage E_{ms} by (7)

$$E_{ms} = 2N_g\phi_s\omega. \qquad (7)$$

We finally obtain

$$E_c = 2(E_m - E_{ms})\frac{R_cN_g}{R_0N_c\pi}. \qquad (8)$$

Thus, as long as E_c does not exceed the value given in (8), α will be independent of E_c.

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Ft. Monmouth, N. J.

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* Received by the IRE, September 14, 1959.

[1] H. F. Storm, "Magnetic Amplifiers," John Wiley and Sons, Inc., New York, N. Y.; 1955.

Harmonic Generation Using Idling Circuits*

Manley and Rowe[1] have treated the case of mixing and harmonic generation by a nonlinear reactance. They show that for the latter case

$$\sum_{m=1}^{\infty} W_{m,0} = 0 \qquad (1)$$

where $W_{m,0}$ is the algebraic power delivered to the nonlinear reactance of frequency mf_1; the fundamental frequency is f_1. In practice, one is interested in converting from frequency f_1 to a harmonic kf_1. It is commonly believed that this is automatically accomplished by terminating all circuits reactively except those of frequencies f_1 and kf_1.

The purpose of this note is to show that:

1) conversion to an arbitrary order of harmonic requires additional criteria on the order of the nonlinearity of the frequency converting element;
2) by terminating additional circuits resistively in such a way as to allow large circulation of energy at additional frequencies, it is possible to increase the order of the harmonic that can be generated over that allowed by the two-circuit system.

The schematic diagram of a harmonic generator using an element with a square-law characteristic is shown in Fig. 1. Here

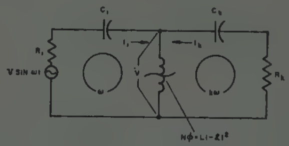

Fig. 1—Two-loop harmonic generator.

the two loops are tuned to resonance at frequencies ω and $k\omega$, respectively. They are coupled by a nonlinear inductance whose characteristic is given by

$$N\phi = Li - \mathcal{L}i^2, \qquad (2)$$

where ϕ and i are the instantaneous deviations of flux and current from those of a static operating point. The voltage across the nonlinear element is then

$$V = L\frac{di}{dt} - 2\mathcal{L}i\frac{di}{dt}, \qquad (3)$$

where the second term on the right is the one of importance for harmonic generation. If the circuit Q's are high (a requirement for efficient harmonic generation),[2] then the currents in the two meshes may be expressed as

$$i_1 = I_1 \sin(\omega t + \theta_1)$$
$$i_k = I_k \sin(k\omega t + \theta_k). \qquad (4)$$

Since $i = i_1 + i_k$, the second term in (3) becomes

$$\begin{aligned}-2\mathcal{L}i\frac{di}{dt} = {} & -I_1{}^2\mathcal{L}\omega \sin(2\omega t + 2\theta_1)\\ & -I_1Ik\mathcal{L}_k\omega \sin[(k+1)\omega t + \theta_1 + \theta_k]\\ & +I_1Ik\mathcal{L}_k\omega \sin[(k-1)\omega t + \theta_k - \theta_1]\\ & -I_1I_k\mathcal{L}\omega \sin[(k+1)\omega t + \theta_k + \theta_1]\\ & -I_1I_k\mathcal{L}\omega \sin[(k-1)\omega t + \theta_k - \theta_1]\\ & -I_k{}^2\mathcal{L}k\omega \sin(2k\omega t + 2\theta_k). \qquad (5)\end{aligned}$$

If power transfer from the "ω" circuit to the "$k\omega$" circuit is to occur, terms of frequency ω and $k\omega$ must exist simultaneously in this expression. The frequencies present are

$$2\omega,\ 2k\omega,\ (k+1)\omega,\ (k-1)\omega.$$

We see that coupling can exist only for two values of k, *viz.*, $k=1$ (trivial) and $k=2$.

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* Received by the IRE, December 18, 1959.

[1] J. M. Manley and H. E. Rowe, "Some general properties of nonlinear elements," PROC. IRE, vol. 44, pp. 904–913; July, 1956.

[2] K. K. N. Chang, "Harmonic generation with non-linear reactances," *RCA Rev.*, vol. 19, pp. 455–464; September, 1958.

Therefore, such a circuit using a square-law element will produce negligible amounts of harmonic outputs of order higher than the second.

If the circuit Q's are low, the current in each loop will not be restricted to only the resonant frequency of that loop. The additional components will mix in the nonlinear reactance to produce power at higher harmonics. This power will be very small.

However, appreciable amounts of power at the third or fourth or even higher harmonics can be achieved by the addition of an idling circuit to the basic two-loop configuration, as shown in Fig. 2. With this scheme, energy is transferred to the 3ω loop as the result of the mixing of current at the fundamental frequency with that produced in a resonant "idler" at 2ω. We have analyzed this circuit as well as the similar ω-2ω-4ω circuit, following the method of Bloom and Chang.[3] The current in the fundamental loop in both cases obeys a fifth-degree algebraic equation. For the 3ω case, the equation is

$$\left(\frac{I_1}{i_0}\right)^5\left[\frac{9}{Q_3}+\frac{36}{Q_1}\right]-\frac{36}{Q_0}\left(\frac{I_1}{i_0}\right)^4 + \left[\frac{48}{Q_1Q_2Q_3}+\frac{4}{Q_2Q_3^2}\right]\left(\frac{I_1}{i_0}\right)^3 - \frac{48}{Q_0Q_2Q_3}\left(\frac{I_1}{i_0}\right)^2+\frac{16}{Q_1Q_2^2Q_3^2}\left(\frac{I_1}{i_0}\right) - \frac{16}{Q_0Q_2^2Q_3^2}=0, \tag{6}$$

where

$$i_0=\frac{L}{2\mathfrak{L}};\ Q_j=\frac{\omega L}{R_j}\ (j=0,1,2,3);$$

$$R_0=V/i_0.$$

For the 4ω case:

$$\left(\frac{I_1}{i_0}\right)^5+\left[\frac{4}{Q_1Q_2}+\frac{16Q_4}{Q_1^3}\right]\left(\frac{I_1}{i_0}\right)^3 - \left[\frac{4}{Q_0Q_2}+\frac{48Q_4}{Q_0Q_1^2}\right]\left(\frac{I_1}{i_0}\right)^2 + \frac{48Q_4}{Q_0^2Q_1}\left(\frac{I_1}{i_0}\right)-\frac{16Q_4}{Q_0^3}=0. \tag{7}$$

Figs. 3 and 4 illustrate representative cases for the third and fourth harmonic conversion, where the efficiency η is defined as

$$\eta=\frac{\text{output power at } k\omega}{\text{total input power}}.$$

In general, these as well as other data show that the efficiency increases with circuit Q's and input voltage. Of course, the efficiency is limited by the maximum allowable current in the element, as would also be true in a simple converter.

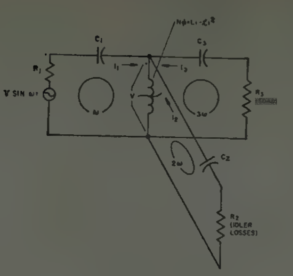

Fig. 2—Harmonic generator with idler.

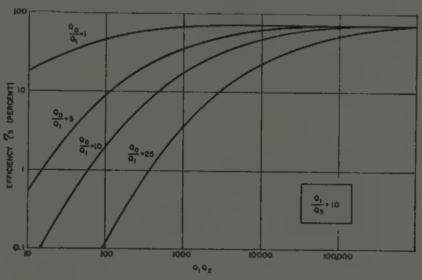

Fig. 3—Efficiency of 3ω conversion vs circuit parameters.

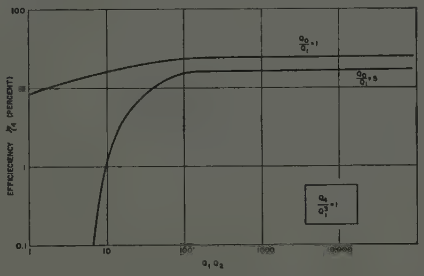

Fig. 4—Efficiency of 4ω conversion vs circuit parameters.

Experimental work which substantiates these results was performed at audio frequencies. To simulate an element which was nearly square law, a toroidal inductor was current-biased to a square-law region. The insertion of this inductor into a two-loop circuit yielded only the second harmonic of the fundamental 1000-cycle signal. When the output loop was tuned to either 3ω or 4ω, no measurable output was present. Operation of the same element in the three-loop circuit (ω, 2ω, 3ω, or ω, 2ω, 4ω) with the same driving voltage yielded considerable amounts of third (or fourth) harmonic power. Table I shows a comparison between the theoretical and experimental results.

TABLE I

Three-loop circuit	Q_0	Q_1	Q_2	Q_3	Q_4	Experimental I_1/i_0	Theoretical I_1/i_0	Experimental η_3	Theoretical η_3	Experimental η_4	Theoretical η_4
ω-2ω-3ω	4.50	2.86	1.44	0.751	—	0.641	0.520	8.5 per cent	7.2 per cent	—	—
ω-2ω-4ω	4.8	2.9	1.5	—	8.86	0.790	0.490	—	—	3.0 per cent	0.57 per cent

The deviations between the theoretical and experimental values are probably due to the following: 1) the assumption that current at $n\omega$ flows only in the $n\omega$ loop was not strictly true because of low Q's; and 2) in the calculation of the efficiencies, it was necessary to raise to a power the difference between two numbers of nearly equal magnitude.

The work reported here has treated only a square-law device. Enhancement of the third harmonic generated by a ferrite at microwave frequencies using this general feedback scheme has recently been reported.[4] For elements that have nonlinearities higher than the second, such as the back-biased diode, efficient conversion to a higher harmonic is of course possible with the two-loop circuit. However, even here the efficiency should be improved by resistively terminating additional branches of the Manley-Rowe circuit, as described above.

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[3] S. Bloom and K. K. N. Chang, "Theory of parametric amplification," *RCA Rev.*, vol. 18, pp. 578–593; December, 1957.

[4] E. N. Skomal and M. A. Medina, "Study of a microwave ferrimagnetic multiple signal conversion process," *J. Appl. Phys.*, vol. 30 (suppl.), pp. 161S–162S; April, 1959.

Solid State Generator for 2×10^{-10} Second Pulses*

Oscillograms showing pulses having indicated base durations as short as 2×10^{-10} seconds have been obtained using an improved version of the diffused silicon mesa diode described by Forster and Zuk.[1]

The diode under test is mounted in shunt of a 50-ohm coaxial line with provision for dc bias as shown in the schematic, Fig. 1. In this application the diode is biased in the forward direction. A 20-mc generator drives the diode with positive pulses of approximately 10^{-8} seconds duration at the base. A sharp step is produced on the trailing edge of the generator pulses by the action of the shunt diode at about the time the diode stops conducting. The differentiating stub

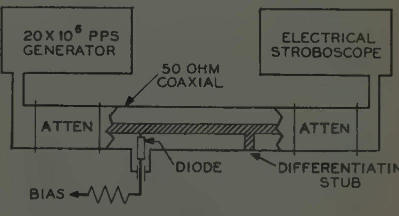

Fig. 1—Block schematic for pulse generator.

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* Received by the IRE, January 13, 1960.

[1] J. H. Forster and P. Zuk, "Millimicrosecond diffused silicon computer diodes," 1958 WESCON CONVENTION RECORD, pt. 3, pp. 122–130.

translates the step voltage into the 2×10^{-10} second pulse shown in the oscillogram of Fig. 2. This pulse has an amplitude of 0.5 volt and a repetition rate of 20 mc.

An example of a sharp voltage step is shown in Fig. 3. In this case the sharp step was produced by an 80-mc sine wave generator instead of the 20-mc pulse generator.

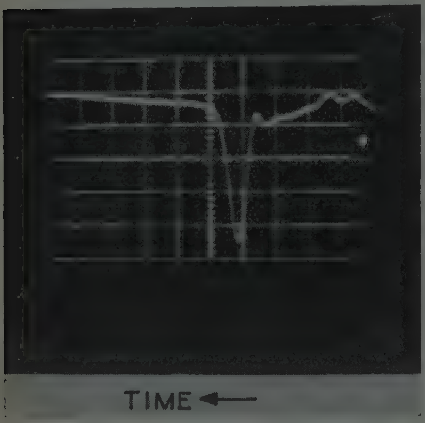

Fig. 2—20-mc pulse, base duration 2×10^{-10} seconds, amplitude 0.5 volt.

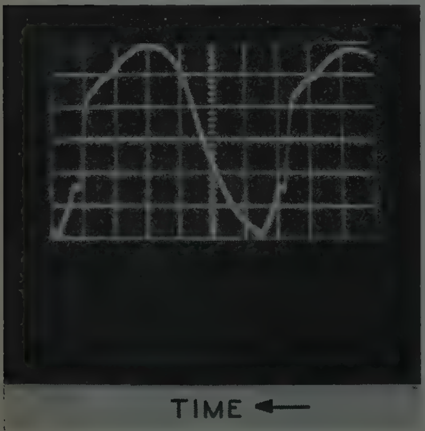

Fig. 3—80-mc sine wave with step, peak to-peak signal 3 volts.

Either the sharp steps or the short pulses can be produced at any repetition rate up to several hundred million per second and there appears to be no limit to the low frequency repetition rate. In other words, a single pulse of approximately 10^{-10} seconds duration can be produced provided that an appropriate single driving pulse is available.

These oscillograms were made using the "fractional millimicrosecond electrical stroboscope" which will be described elsewhere.[2]

We wish to thank J. H. Forster of the Bell Telephone Laboratories for supplying us with the special computing diodes used in these experiments.

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W. M. Goodall
Bell Telephone Labs., Inc.
Holmdel, N. J.

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[2] W. M. Goodall and A. F. Dietrich, "Fractional Millimicrosecond Electric Stroboscope," submitted to Proc. IRE.

Extension of Longitudinal-Beam Parametric-Amplifier Theory*

The longitudinal-beam parametric amplifier has been previously investigated[1,2] using a model in which coupling between only the lower sideband and the signal was considered. Experiments by Ashkin, *et al.*[3] which showed large discrepancies between theoretical and measured values of gain, raised the question as to whether or not the original model accurately described the system. Another model which had been used in study of high-frequency beam modulations of traveling-wave tubes is now proposed. In the latter model, coupling is allowed between the signal and the first upper and lower sidebands around the fundamental of the pump frequency. The coupling to the sidebands around the second harmonic of the pump may be an important factor; however, this is neglected in the present study.

The assumptions made in this "multimode" theory are very similar to those used by Louisell and Quate;[1] that is, the pump frequency is twice the signal frequency, only one space-charge wave (either fast *or* slow) is excited at each frequency, and the pump amplitude is much greater than the sideband and signal amplitudes. All amplitudes are assumed much less than the beam average quantities. Under these conditions the pump signal will propagate unperturbed with a wave number

$$\beta_e\left(1-\frac{a'\omega_q}{\omega}\right),$$

where

$$a'=\frac{\omega_{q2}}{2\omega_q}=\frac{\text{reduced plasma frequency at pump frequency}}{2x\text{ reduced plasma frequency at signal frequency}},\qquad(1)$$

and $a'>0$ for the fast wave and $a'<0$ for the slow wave. Introduce the following definitions:

$$b'=\frac{\omega_{q3}}{3\omega_q}=\frac{\text{reduced plasma frequency at upper sideband}}{3x\text{ reduced plasma frequency at signal frequency}}\qquad(2)$$

$$m=|m|e^{j\phi}=\frac{i}{I_0}=\begin{array}{l}\text{complex modulation}\\ \text{index of pump}\\ \text{excitation,}\end{array}\qquad(3)$$

where i is component of RF convection current at the pump frequency and I_0 is the average beam current. Also v_1 is defined as the velocity at the signal frequency, v_{11} as the velocity at the upper sideband and, finally, $\beta_q=\omega_q/u_0$ as the reduced plasma frequency radian wave number. Assume that the velocities can be described by

$$v_1=u_1(z)\exp-j\beta_e\left(1-\frac{a'\omega_q}{\omega}\right)z,$$

$$v_{11}=u_{11}(z)\exp-3j\beta_e\left(1-\frac{a'\omega_q}{\omega}\right)z.\qquad(4)$$

The propagation at the sideband frequencies can then be described in terms of the velocity amplitudes by two simultaneous linear differential equations:

$$\begin{aligned}&\frac{d^2u_1}{dz^2}+2j\beta_qa'\frac{du_1}{dz}\\ &\quad-\beta_q^2\left[a'^2\left(1+\frac{|m|^2}{4}\right)\right.\\ &\quad\left.-1+\frac{a'^2}{b'^2}\frac{|m|^2}{4}\right]u_1\\ &\quad-\beta_q^2\frac{m}{2}(1+2a'^2)u_1^*\\ &\quad+j\beta_qa'\frac{m^*}{2}\left[1+\frac{1}{3b'^2}\right]\frac{du_{11}}{dz}\\ &\quad-\beta_q^2\frac{m^*}{2}\left[1+a'^2\left(1+\frac{1}{b'^2}\right)\right]u_{11}\\ &\quad-a'^2\beta_q^2\frac{m^2}{4}u_{11}^*=0\end{aligned}\qquad(5)$$

and

$$\begin{aligned}&\frac{d^2u_{11}}{dz}+6j\beta_qa'\frac{du_{11}}{dz}\\ &\quad-9\beta_q^2\left[a'^2(1+b'^2)\frac{|m|^2}{4}-b'^2\right]u_{11}\\ &\quad+j\beta_qa'\frac{3m}{2}[1+3b^2]\frac{du_1}{dz}\\ &\quad-9\beta_q^2\frac{m}{2}[a'^2(1+b'^2)+b'^2]u_1\\ &\quad-9a'^2b'^2\beta_q^2\frac{m^2}{4}u_1^*=0.\end{aligned}\qquad(6)$$

The asterisk indicates the complex conjugate of a quantity and the assumption $(a'\omega_q/\omega)\ll1$ has been made. Eqs. (5) and (6) can be integrated by standard linear methods using the operator $d/dz=\beta_q\mu$. There will be eight waves propagating in this system. An approximate solution for the case of a thin beam $(a'=b'=1)$ and for small pump amplitudes, $|m|<1$, is given as

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* Received by the IRE, January 7, 1960.

[1] W. H. Louisell and C. F. Quate, "Parametric amplification of space-charge waves," Proc. IRE, vol. 46, pp. 707–716; April, 1958.

[2] H. A. Haus, "The kinetic power theorem for parametric longitudinal beam amplifiers," IRE Trans. on Electron Devices, vol. ED-5, pp. 225–232; October, 1958.

[3] A. Ashkin, T. J. Bridges, W. H. Louisell, and C. F. Quate, "Parametric electron beam amplifiers," 1958 WESCON Convention Record, pt. 3, pp. 13–22.

$$\mu_{1,2} \approx \pm j6.0$$
$$\mu_{3,4} \approx \pm j2.0$$
$$\mu_{5,6} = 0.375\,|m| \pm j1.25\,|m|$$
$$\mu_{7,8} \approx -0.375\,|m| \pm j1.25\,|m|. \tag{7}$$

More exact values of the propagation constants are shown in Figs. 1 and 2 respectively for a thin beam and a beam of finite thickness. The wave amplitudes can be calculated by using the boundary conditions that the signal is initially excited in the beam as a pure single space-charge wave and that the initial upper sideband velocity is zero. It can then be shown that only the growing or declining waves are appreciably excited. Gain can be obtained by adjusting the pump phase to $\Phi = \pi/2$. The velocity wave amplitudes for a thin beam are then described as

$$u_1 \approx 1.05 \cos\,(1.25\,|m|\,\beta_q z - 17.84°)$$
$$\exp 0.375\,|m|\,\beta_q z \exp - j\beta_e z \left(1 - \frac{a'\omega_q}{\omega}\right) \tag{8}$$
$$u_{11} \approx 1.822 \sin\,(1.25\,|m|\,\beta_q z) \exp 0.375\,|m|\,\beta_q z$$
$$\exp - 3j\beta_e z \left(1 - \frac{a'\omega_q}{\omega}\right) \cdot \tag{9}$$

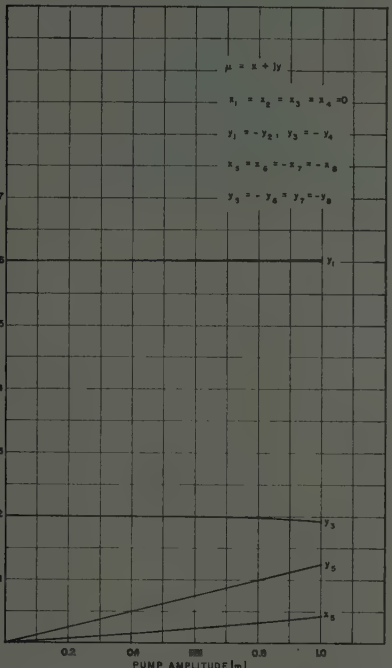

Fig. 1—Longitudinal-beam parametric-amplifier propagation constants as functions of pump amplitude $a' = b' = 1$.

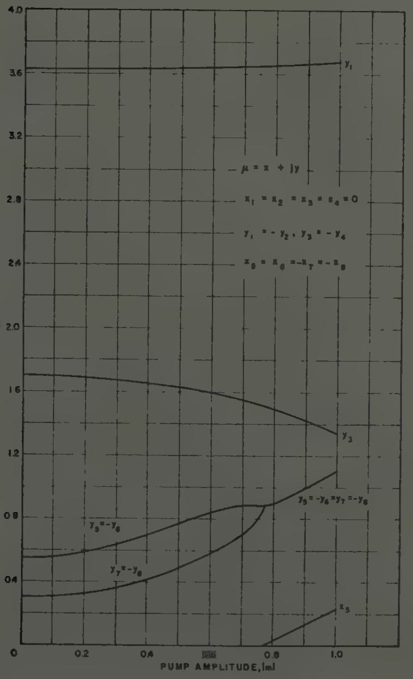

Fig. 2—Longitudinal-beam parametric-amplifier propagation constants as functions of pump amplitude $a' = 0.697$, $b' = 0.513$.

It can be concluded that the upper sideband is heavily excited in the beam. The growth constant is one half that obtained by Louisell and Quate;[1] therefore, the theoretical gain will be much lower than theirs and hence closer to the experimental value. The next refinement in the analysis would be the inclusion of the sidebands around the second harmonic of the pump. It is expected that this will provide still better agreement between theory and experiment, since the growth rate will be further reduced.

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WWV Standard Frequency Transmissions*

The frequencies of the National Bureau of Standards radio stations WWV and WWVH are kept in agreement with respect to each other and have been maintained as constant as possible with respect to an improved United States Frequency Standard (USFS) since December 1, 1957.

The nominal broadcast frequencies should, for the purpose of highly accurate scientific measurements, or of establishing high uniformity among frequencies, or for removing unavoidable variations in the broadcast frequencies, be corrected to the value of the USFS, as indicated in the table below.

The characteristics of the USFS, and its relation to time scales such as ET and UT2, have been described in a previous issue,[1] to which the reader is referred for a complete discussion.

The WWV and WWVH time signals are also kept in agreement with each other. In addition, they are locked to the nominal frequency of the transmissions and consequently may depart continuously from UT2. Corrections are determined and published by the U. S. Naval Observatory. The broadcast signals are maintained in close agreement with UT2 by properly offsetting the broadcast frequency from the USFS at the beginning of each year when necessary. This new system was begun on January 1, 1960. The last time adjustment was a retardation adjustment of 0.02 seconds on December 16, 1959.

WWV Frequency
With Respect to U. S. Frequency Standard

1960 January 1600 UT	Parts in 10^{10}†
1	−149
2	−149
3	−149
4	−149
5	−148
6	−148
7	−148
8	−148
9	−148
10	−148
11	−148
12	−148
13	−148
14	−147
15	−147
16	−146
17	−146
18	−145
19	−145
20	−145
21	−145
22	−145
23	−145
24	−145
25	−145
26	−145
27	−145
28	−145
29	−145
30	−145
31	−145

† A minus sign indicates that the broadcast frequency was low.

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Tunnel Diode as an Interstage Gain Device*

The purpose of this letter is to demonstrate the applicability of the tunnel diode as an interstage coupling element with gain. It can be used in a transistor amplifier consisting of a cascade of "common-base" stages to provide impedance transformation and power gain between each stage. Fig. 1 illustrates the configuration used to obtain this impedance transformation and power gain. The gain equations are as follows:

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* Received by the IRE, February 23, 1960.

[1] "United States national standards of time and frequency," Proc. IRE, vol. 48, pp. 105–106; January, 1960.

* Received by the IRE, December 17, 1959.

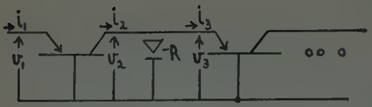

Fig. 1—Circuit configuration for tunnel diode interstage coupling, where $u_2=u_3$, and for $\infty_6=1$, $i_1=i_2$ (valid assumption for most transistors).

$$G_T=\frac{P_2}{P_1}=\frac{v_2i_2}{v_1i_1}=\frac{(-R)}{R_{\text{in}}+(-R)}$$

$$G_D=\frac{P_3}{P_2}=\frac{v_3i_3}{v_2i_2}=\frac{(-R)}{R_{\text{in}}+(-R)}$$

$$G_S=\frac{P_3}{P_1}=\frac{v_3i_3}{v_1i_1}=G_TG_D=\left[\frac{(-R)}{R_{\text{in}}+(-R)}\right]^2$$

$$\text{for stable gain } |R_{\text{in}}| < |-R|$$

$(-R)$ = negative resistance of tunnel diode
R_{in} = input resistance of common base transistor
G_t = transistor power gain
G_d = power gain per diode (interstage gain)
G_s = power gain per stage.

Since presently available diodes have negative resistance values of −40 to −200 ohms and since R_{in} for typical transistors can be controlled in this range, it is fairly simple to obtain the gain and stability desired.

A three stage IF amplifier using this coupling technique was designed and breadboarded. An over-all gain of 60 db was obtained for a bandwidth of 300 kc centered at 10 mc. The gain per transistor and the gain per diode were equal to 10 db each. Experimental tunnel (Esaki) diodes from the Sony Company in Japan were used; these had a negative conductance of −0.02 mho and a shunt capacitance of 500 $\mu\mu$f when operated in the tunnel region. The large value of shunt capacitance (500 $\mu\mu$f) was the limiting factor in the determination of center frequency and bandwidth for the desired gain. However, tunnel diodes are now being manufactured which have shunt capacitances of 15 $\mu\mu$f with the expectation that this will be reduced further in the future.

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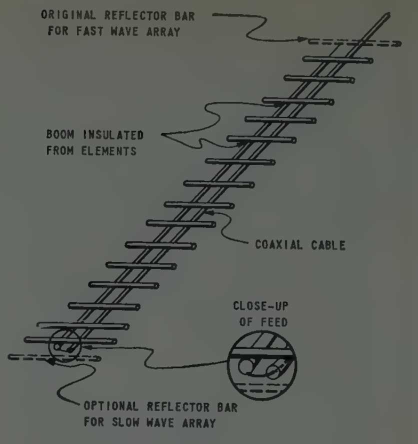
Recent Developments in Very Broad-band End-Fire Arrays*

The transmission line excited Yagi or "straight ladder" antenna, shown in Fig. 1, was developed at this laboratory in the spring of 1958. This antenna may be described as a conventional Yagi array in which the driving element has been replaced by a shunted passive element and the array excited by electromagnetic coupling to a twin-wire transmission line, the twin-wire line being developed from a coaxial line in the manner shown in Fig. 1. In Fig. 2 we show a sketch of an experimental model of a two-bay straight ladder. This antenna was ultimately built full scale for operation at VHF (Fig. 3). It had a 20-foot boom length, 50-ohm input impedance, and a minimum over-all gain of 10 db (above isotropic) in a bandwidth of ±10 per cent.



ORIGINAL REFLECTOR BAR FOR FAST WAVE ARRAY
BOOM INSULATED FROM ELEMENTS
COAXIAL CABLE
CLOSE-UP OF FEED
OPTIONAL REFLECTOR BAR FOR SLOW WAVE ARRAY

Fig. 1—Sketch of a straight ladder antenna.

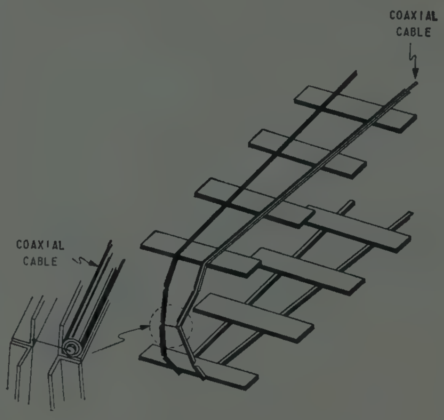


Fig. 2—Two-bay straight ladder antenna.

It was subsequently realized that the ladder could be made indefinitely broad in bandwidth by using the principle employed by DuHamel and Isbell,[1] *i.e.*, by scaling the radiating elements in constant ratio. A sketch of one of our first working models of this antenna, called the "Tapered Ladder," is shown in Fig. 4. As in its predecessor, the straight ladder, the elements were electromagnetically coupled, *i.e.*, there were no conducting connections between the twin-line and the elements.

Typical *E*- and *H*-plane radiation patterns of a Tapered Ladder are shown in Fig. 5. These patterns were maintained over a bandwidth exceeding 4:1. The bandwidth can be increased simply by adding more elements. The antenna was excited from 50-ohm coaxial line, had a maximum VSWR of 2:1, and a measured gain of 8 db above isotropic.

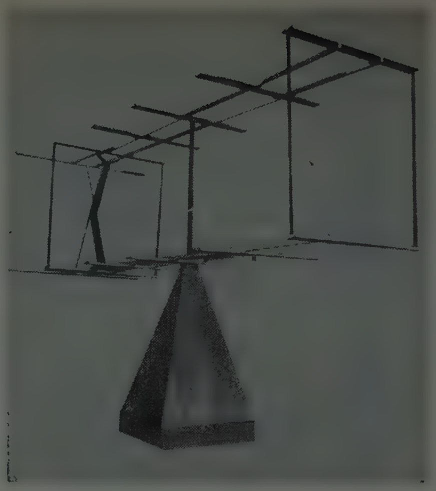


Fig. 3—Photograph of a two-bay VHF straight ladder antenna.

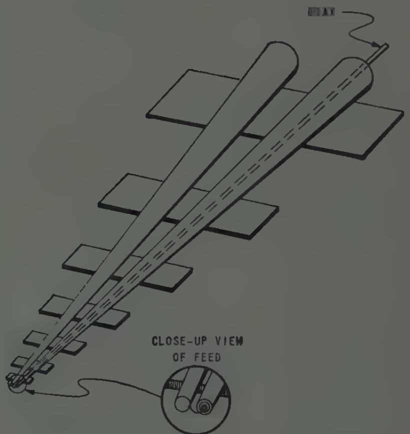


Fig. 4—Sketch of an early working model of a tapered ladder antenna.

In a later version of the Tapered Ladder we were able to match the antenna to 50 ohms with an average VSWR of 1.45:1 over a 7:1 frequency range. This was achieved by providing additional distributed loading for the twin-conductor transmission line. We have also built models with higher directivities, increase in directivity being derived from a reduction of the taper angle of the array. The taper angle is that angle subtended at the antenna vertex by each radiating element; thus, reduction of the taper angle amounts to reduction of percentage change in length of successive elements.

Other types of end-fire antennas can be made very broadband by tapering and exciting from twin-line, in particular the "Helix" and Cumming's "Zig Zag."[2] A broadband "Tapered Zig-Zag" was developed by J. McDowell and the author. A sketch of such an antenna is shown in Fig. 6 and typical radiation patterns are shown in Fig. 7. The radiating structure of a Tapered Zig-Zag can be a slightly modified form of one-half of the V-shaped undirectional, pseudo-infinite, logarithmically periodic

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* Received by the IRE, August 20, 1959.

[1] R. H. DuHamel and D. E. Isbell, "Broadband logarithmically periodic antenna structures," 1957 IRE National Convention Record, pp. 119–128.

[2] W. A. Cumming, "A nonresonant endfire array for VHF and UHF," IRE Trans. on Antennas and Propagation, vol. AP 3, pp. 52–58; April, 1955.

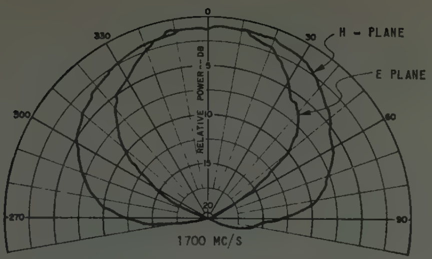

Fig. 5—Typical E- and H-plane radiation patterns of an early model tapered ladder antenna.

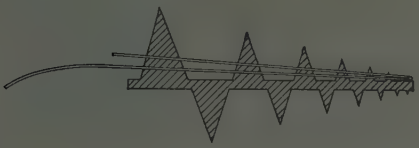

Fig. 6—Broadband tapered zig-zag.

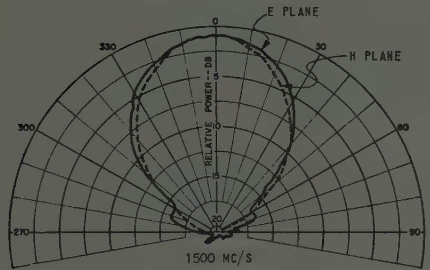

Fig. 7—Broadband tapered zig-zag radiation patterns, E- and H-planes at 1500 mc.

(PILP) antenna developed by Isbell.[3] In such case it is remarkable to note that the zig-zag structure has the same directivity as the original PILP.

We have found that the Tapered Ladder and the symmetrically excited Tapered Zig-Zag can be cut in half longitudinally and excited over a ground plane. Further, if the radiating elements are rotated 90° about their own longitudinal axes we have a transmutation to the broadband "Loaded Unipole" independently developed by Lamberty[4] at this laboratory.

The author and co-workers have made near-field amplitude measurements of the Tapered Ladder, and Franks and Bell[5] have made near-field phase and amplitude measurements of a PILP structure. In both cases we find that at any given frequency the antennas derive their directivity from an excitation length only half that of an equivalent Hansen-Woodyard array. Further, Franks and Bell find that the excess phase shift along the boom exceeds the 180° Hansen-Woodyard condition, the measured value being greater than 250°.

If we regard PILP structures, ladders, apex-fed helices[6] and spirals, line-fed zig-zags, etc. as loaded transmission lines, all of them appear to be generically related and resemble the "Comb" and "Fishbone" antennas developed several decades ago. With an analysis based on the loaded transmission line point of view, we have had fair success in predicting the radiation patterns of the Tapered Ladder. Relationships between excitation length, bandwidth, and directivity have been deduced from the uncertainty principle.

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[3] D. E. Isbell, "Non-planar logarithmically periodic antenna structures," Seventh Annual Symposium on the USAF Antenna Res. and Dev. Program, University of Illinois, Urbana; October 21–25, 1957.

[4] B. J. Lamberty, "A class of low gain broadband antennas," 1958 IRE WESCON Convention Record, pp. 251–259.

[5] R. Franks and R. Bell, Sylvania Electronic Defence Lab., Mountain View, Calif., private communication.

[6] H. S. Barsky, "Broadband conical helix antennas," 1959 IRE National Convention Record, pp. 138–146.

Nonreciprocal Radar Antennas*

We herewith report a successful extrapolation of the Dolph-Tchebycheff antenna concept with application to radar antennas or to any situation in which the system antenna pattern is the product of a transmitting antenna pattern and the pattern of a dimensionally-equivalent receiving antenna.

We begin by pointing out the inadequacy of the simple Dolph-Tchebycheff design in the radar case. In this design, the array factor for $(n+1)$ elements is forced by appropriate choice of feeding coefficients into a correspondence with T_n, where T_n denotes the Tchebycheff polynomial of degree n. Dolph[1] pointed out that the Tchebycheff correspondence produces the minimum beam angle for a given minor-lobe level. However, for the radar case in which the same antenna, or identical antennas, are used for transmission and reception, the system pattern (pattern and array factor are taken as equal) is $(T_n)^2$. $(T_n)^2$ is a polynomial of degree $2n$ but it is *not* equal to the Tchebycheff polynomial of the same degree (T_{2n}) and therefore the system pattern does *not* represent the optimum relationship between beam angle and minor-lobe level.

It is, however, quite simple to produce a system pattern with the desired Tchebycheff correspondence. Note first that T_{2n} is expressible in terms of T_n in the same way that the cosine of a double angle is expressible in terms of the cosine of the single angle. Thus

$$T_{2n} = 2(T_n)^2 - 1,$$

or

$$T_{2n} = 2\left(T_n - \frac{\sqrt{2}}{2}\right)\left(T_n + \frac{\sqrt{2}}{2}\right).$$

Therefore, if the excitation of the array center is adjusted to the different values between transmission and reception as implied by the above equation, the Tchebycheff correspondence for system pattern will have been realized.

This nonreciprocal action can be produced in a number of ways, perhaps the most simple of which is shown in Fig. 1. Here the power divider apportions the excitation of the array elements in accordance with $(T_n+\sqrt{2}/2)$. The isolator action is such as to leave this excitation unaltered for one direction of power flow and to introduce enough attenuation in the reverse direction to correspond to the $(T_n-\sqrt{2}/2)$ excitation.

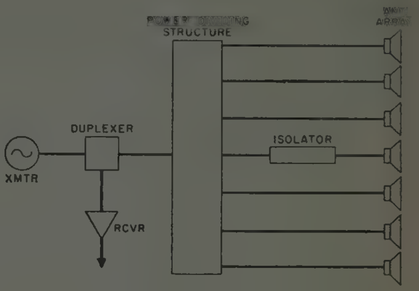

Fig. 1—Extended Tchebycheff nonreciprocal array.

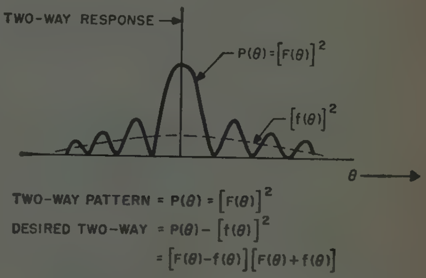
TWO-WAY RESPONSE
P(θ)=[F(θ)]²
[f(θ)]²
θ
TWO-WAY PATTERN = P(θ) = [F(θ)]²
DESIRED TWO-WAY = P(θ) − [f(θ)]²
= [F(θ)−f(θ)][F(θ)+f(θ)]

Fig. 2—Auxiliary radiator used to simultaneously sharpen beam angle and reduce sidelobes of two-way antenna pattern.

Using a given minor-lobe level as the basis of comparison, the extended Tchebycheff design will improve the beam angle by about 10 per cent over the conventional Tchebycheff design. For a given beam angle, the equivalent one-way minor-lobe improvement lies in the 4-to 5-db range.

This general principle has the simple continuous aperture analog illustrated in Fig. 2. Here the usual two-way pattern is shown as $[F(\theta)]^2$. The dotted $[f(\theta)^2]$ curve is selected to bisect exactly the largest minor lobe and may also approximately bisect several of the lesser minor lobes as well. If $[f(\theta)]^2$ could be made the angular axis (abscissa) then the minor-lobe level would be reduced by nearly 3 db and the beam angle slightly sharpened at the same time. It may be obvious by now that this effect can be produced by an auxiliary antenna whose one-way pattern is $f(\theta)$. This auxiliary antenna adds in phase to the principal lobe of the main antenna for transmission or reception, and adds in phase opposition to the principal lobe for reception or transmission. This action may be described by the algebraic relation:

$$[F(\theta)]^2 - [f(\theta)]^2 = [F(\theta) - f(\theta)][F(\theta) + f(\theta)].$$

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* Received by the IRE, August 20, 1959. The work reported herein was sponsored by the Ordnance Corps in connection with Contract DA-30-069-ORD-1955.

[1] C. L. Dolph, "A current distribution for broadside arrays which optimizes the relationship between beamwidth and sidelobe level," Proc. IRE, vol. 34, pp. 335–345; June, 1946.

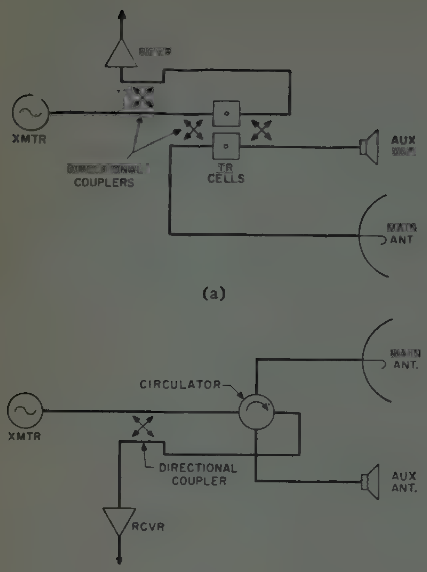

Fig. 3—Mechanisms for addition and subtraction of auxiliary radiator.

Possible mechanisms for this alternate addition and subtraction are shown in Fig. 3.

The ideas described in the foregoing have been verified experimentally for both arrays and continuous apertures at these laboratories.

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Note on the Noise Figure of Negative Conductance Amplifiers*

The problem of coupling a negative conductance amplifier stage to another stage is sometimes solved by means of a circulator.[1] The purpose of this note is to discuss two solutions that do not require the use of a circulator. One solution consists of connecting the negative conductance amplifier stage directly to a receiver; the other solution connects the stage to a receiver by means of a lossless step-down transformer of suitably chosen turn ratio n.

Let the negative conductance amplifier stage consist of a negative conductance device of conductance $-g_d$ parallel to a signal source of conductance g_s, where $g_s > g_d$. The available gain of the stage is then (see Fig. 1):

$$G_{av} = g_s/(g_s - g_d), \tag{1}$$

and the amplifier stage is stable. Let the noise of the amplifier be represented by a source current generator i_s and a device current generator i_d and let

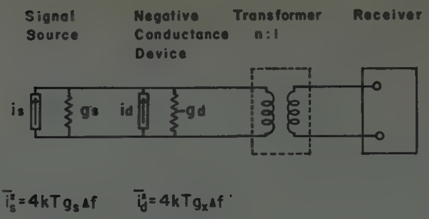

Fig. 1.

$$\overline{i_s^2} = 4kTg_s\Delta f; \quad \overline{i_d^2} = 4kTg_x\Delta f. \tag{2}$$

The noise figure F_1 of the amplifier stage, neglecting losses, is then

$$F_1 = 1 + g_x/g_s. \tag{3}$$

Let the receiver have a noise figure $F_2(g_s')$ if a signal source of conductance g_s' is connected parallel to the input, and let F_2 have the minimum value $(F_2)_{\min}$ for $g_s' = (g_s')_{\min}$. If the negative conductance amplifier stage is connected directly to the input of the receiver, the receiver looks into a source conductance $g_s' = (g_s - g_d)$; if it is connected by means of a step-down transformer, the receiver looks into a source conductance $g_s' = n^2(g_s - g_d)$, if n is the turn ratio of the transformer. In either case $g_s' > 0$, since $g_s > g_d$.

Since the output conductance $(g_s - g_d)$ of the negative conductance amplifier stage is positive, Friis' formula holds. The total noise figure F of amplifier plus receiver is

$$F = F_1 + (F_2 - 1)/G_{av}. \tag{4}$$

We first consider the case in which the negative conductance amplifier is connected directly to the receiver and we assume that $g_s' = (g_s - g_d)$ is quite small. The asymptotic value of the noise figure is then

$$F = 1 + \left[g_x + \lim_{g_s' \to 0} g_s'\{F_2(g_s') - 1\} \right] / g_s, \tag{5}$$

as is found by substituting (1). Hence, the direct connection between the stage and the receiver does not give a significant increase in noise figure if

$$\lim_{g_s' \to 0} g_s'\{F_2(g_s') - 1\} \ll g_x. \tag{5a}$$

Next, we consider the case in which a step-down transformer of turn ratio n is used to connect the stage to the receiver. The minimum noise figure $F_{\min}$ is obtained if the source conductance $g_s' = n^2(g_s - g_d)$ is adjusted so that F_2 has its minimum value. Substituting (1), we obtain for $n^2(g_s - g_d) = (g_s')_{\min}$,

$$F = F_{\min} = 1 + [g_x + (g_s - g_d) \cdot \{(F_2)_{\min} - 1\}]/g_s. \tag{6}$$

Hence, the transformer connection between the stage does not give any significant increase in noise figure if

$$(g_s - g_d)\{(F_2)_{\min} - 1\} \ll g_x. \tag{6a}$$

It is always possible to write the noise figure of the receiver as

$$F_2 = 1 + A + R_{n0}g_s' + g_{n\infty}/g_s', \tag{7}$$

where A is usually a small factor, R_{n0} is the equivalent noise resistance for short-circuited input, and $g_{n\infty}$ is the equivalent noise conductance for open-circuited input; both quantities are easily measured. The noise figure has its minimum value

$$(F_2)_{\min} = 1 + A + 2\sqrt{R_{n0}g_{n\infty}} \quad \text{for } g_s' = (g_s')_{\min} = \sqrt{g_{n\infty}/R_{n0}}. \tag{7a}$$

Consequently (5) may be written

$$F = 1 + (g_x + g_{n\infty})/g_s, \tag{8}$$

and (5a) becomes

$$g_{n\infty} \ll g_x. \tag{8a}$$

Eq. (6) may be written

$$F_{\min} = 1 + \{g_x + (g_s - g_d)(A + 2\sqrt{R_{n0}g_{n\infty}})\}/g_s, \tag{9}$$

and (6a) becomes

$$(g_s - g_d)(A + 2\sqrt{R_{n0}g_{n\infty}}) \ll g_x. \tag{9a}$$

The above discussion is true for any type of negative resistance amplifier such as the parametric amplifier, the tunnel diode, etc. For the tunnel diode the dc current I_d shows full shot noise[2] so that

$$\overline{i_d^2} = 2eI_d\Delta f, \text{ or } g_x = (e/2kT)I_d. \tag{10}$$

In the case of the parametric amplifier, neglecting losses, the noise figure F_1 is[3]

$$F_1 = 1 + \frac{g_{ci}}{g_s}\frac{\omega_i}{\omega_0}, \tag{11}$$

where $-g_{ci}$ is the negative input conductance caused by the idler circuit, ω_i is the input frequency, and ω_0 is the idler frequency. Consequently,

$$g_x = g_{ci}\frac{\omega_i}{\omega_0}. \tag{11a}$$

The discussion has neglected to mention losses in the circuits, in the transformer, and, in the case of the parametric amplifier, the losses in the parametric device itself. It has also ignored the possibility of cooling some of the components, *e.g.*, the parametric device and the idler circuit in parametric amplifiers. It is not difficult to extend the discussion to include these effects.

A tunnel diode has a much lower impedance level than a parametric amplifier and as a consequence (5a) and (6a) and their equivalents are much more easily satisfied for the former than for the latter. Also, the influence of circuit losses will be less pronounced.

The authors are indebted to W. F. Leverton, Raytheon Manufacturing Company, Waltham, Mass., and R. L. Watters, General Electric Research Laboratory, Schenectady, N. Y., for discussion and correspondence.

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* Received by the IRE, September 18, 1959.

[1] S. Bloom and K. K. N. Chang, "Theory of parametric amplification using nonlinear reactances," *RCA Rev.*, vol. 18. pp. 578–593; December, 1957.

[2] J. J. Tiemann and R. L. Watters, "Noise Considerations of Tunnel Diode Amplifiers," Solid-State Device Research Conference, Cornell Univ., Ithaca, N. Y.; June 17–19, 1959.

[3] A. van der Ziel, "Noise figure of reactance converters and parametric amplifiers," *J. Appl. Phys.*, vol. 30; September, 1959.

Contributors

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From 1924 to 1929, he did research and development work for Siemens & Halske A.G., Berlin, on multiple carrier, picture telegraphy over loaded cables, and acoustical standards.

From 1929 to 1935, he was with International Telephone and Telegraph Corp. in London, Paris and Berlin. He was Chief Engineer of the Berlin associates of ITT and did engineering and research in all fields of communications, including magnetic sound recording, wired wireless, common frequency broadcasting.

From 1935 to 1957, he was associated with Allgemeine Electricitaets-Gesellschaft in Berlin and in Frankfurt, Germany. He was Technical Director concerned with the supervision of research and development activities and of patent policy of the concern including associated companies (Telefunken and Osram). He also acted as consultant to the President of the Board and during the war he inaugurated the production of an infrared proximity fuze for the V-2.

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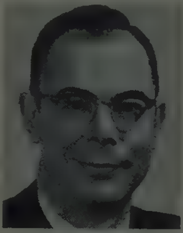
Donald L. Birx (S'49–A'51–M'54) was born in Baltimore, Md., on November 30, 1924. He received the B.E.E. degree from The Johns Hopkins University, Baltimore, Md., in 1944. In 1945, he entered the U. S. Navy and was employed as an instructor in the Radio Materiel School, Washington, D. C. In 1946, he re-entered The Johns Hopkins University in the graduate school, and received the Dr. Eng. degree in electrical engineering in 1951.


D. L. Birx

Since 1951, he has been with the Franklin Institute in Philadelphia, Pa., where he is currently a senior staff engineer in the electronics branch of the laboratories. At present, he is working in the measurements field, being concerned primarily with the measurement of low-level RF powers and the utilization of natural radiation in the navigation of space vehicles. Prior to this work, he was engaged in a variety of projects which have included the statistical design of experiments for the analysis of optical and electrical systems, the synthesis of networks utilized in the design of RF noise filters, and the study of nonlinear aspects of rate gyroscopes.

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Robert L. Blanchard (S'55) was born in Montreal, Can., on May 8, 1919. He received the B.S. degree in engineering physics from McGill University, Montreal, in 1941 and the M.S. degree in applied physics from Harvard University, Cambridge, Mass., in 1948.

R. L. Blanchard

In 1941 and 1942, he was appointed Instructor in physics at McGill University, and in 1953 and 1954 he served as lecturer on microwave theory and technique in the Graduate School, Evening Division, at Northeastern University Boston, Mass. He was in charge of Communication Research and Development for the Canadian Army Signal Corps in 1945 and 1946, following which he joined Raytheon Manufacturing Company, Waltham, Mass., where he developed UHF communication equipment. He was a consultant on electronic development and design for several companies in the Boston area during 1948–1949. In 1949, he joined Trans-Sonics, Inc., Burlington, Mass. where his work has comprised general applied physics with application to measurement and control, electromagnetic theory, microwave systems, network theory, and filter design. He was leader of the team that developed the first successful trajectory scoring system for ammunition and missiles directed at airborne targets. He has held senior positions at Trans-Sonics, Inc., including Chief Research Engineer 1952–1957, Chief Engineer, Electronics Division 1957–1958, and is presently Vice-President and Director of New Product Research.

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Frederick E. Bond (M'47–SM'55) was born on January 10, 1920, in Philadelphia, Pa. He received the B.S.E.E. degree from Drexel Institute of Technology, Philadelphia, in 1941, the M.S.E.E. degree from Rutgers University, New Brunswick, N. J., in 1950, and the D.E.E. degree from Polytechnic Institute of Brooklyn, N. Y., in 1956.

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From 1949 to 1956, he served as instructor and later as assistant professor in the Electrical Engineering Department of Syracuse University, where he was engaged in research work in information theory and microwave antennas and in studies on systems engineering. From 1952 to 1953, on a leave of absence, he was employed in the System Planning Department of the Niagara Mohawk Power Corporation, Buffalo, N. Y., where he was concerned with system planning and econonic operation of a large integrated power system. At present, he is a member of the senior staff of Ramo-Woolridge, a division of Thompson Ramo Wooldridge, Inc., Los Angeles, Calif., where he is working on systems analysis and synthesis, with emphasis on applications of information theory in the field of digital communications. He has also investigated techniques for electronic countermeasures and methods of achieving reliable transmission over fluctuating circuits.

Dr. Cahn is a member of Sigma Xi and the AIEE.

❖

William O. Campbell (M'55) was born on August 26, 1920 in Natchez, Miss. He received the B.S.E.E. and M.S.E.E. degrees in 1948 and 1949, respectively, from the University of Tennessee, Knoxville.

From 1943 to 1946, he served with the U. S. Army Air Corps as a meteorologist. From 1949 to 1955, he was associated with Bendix-Friez in Baltimore, Md., where he was instrumental in the design of an automatic readout mercury barometer and an analog-digital computer for radiosonde data computation. Since 1955, he has been with The Martin Company, Baltimore, where, from 1955 to 1957, he was in charge of design groups for automatic checkout equipment for the Mace and Lacrosse Missiles. During 1957 and 1958, his position as staff Unit Head involved work on the Titan countdown equipment. Presently in the Space Flight Division, he is responsible for the electronic support equipment for space programs. He is currently enrolled in the graduate curriculum of Engineering Management at Drexel Institute of Technology, Philadelphia, Pa.

Mr. Campbell is currently serving on the Test Equipment Committee of the Electronic Industries Association. He has recently filed a patent disclosure for digital checkout equipment.

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R. L. Choate

Robert L. Choate (S'50–A'52–M'57) was born on November 30, 1918, in Reevesville, Ill. He attended the University of Missouri School of Mines and Metallurgy, Rolla, where he received the B.S.E.E. degree in 1950 and the M.S. degree in physics in 1951.

From 1951 to 1953, he was employed by Bell Aircraft Corporation as electronics engineer in the development of missile guidance systems. In 1953, he returned to the Missouri School of Mines, where he served as instructor in physics and engaged in graduate studies in mathematics, physics, and electrical engineering. He has been employed at the Jet Propulsion Laboratory, California Institute of Technology, Pasadena, since 1955 as systems analyst in the Electronics Research Section. There he has been concerned with communication systems analysis, noise and filter theory, phase-locked-loop theory, secure communications, and deep space communications and tracking. He has participated in the Explorer series of satellite experiments and the lunar probe experiments (Pioneer III and IV).

Mr. Choate is a member of Sigma Xi and Sigma Pi Sigma.

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G. M. Clemence was born in Greenville, R. I., on August 16, 1908. He received the Ph.B. degree from Brown University, Providence, R. I., in 1930. He also received an Honorary D.Sc. degree from Case Institute of Technology, Cleveland, Ohio, in 1954.

He has been an astronomer at the U. S. Naval Observatory, Washington, D. C. since 1930. From 1945 to 1958, he was Director of the Nautical Almanac, and since 1958 he has been Scientific Director. He was a visiting professor of astronomy at Columbia University, New York, N. Y., in 1958, and has also written articles on positional astronomy, celestial mechanics time measurement, astronomical constants, mathematical tables, and navigation.

G. M. Clemence

Dr. Clemence is a member of the Senior Scientists Council of the Navy Department, the National Academy of Sciences, the American Academy of Arts and Sciences, the American Astronomical Society, the Bureau des Longitudes (France), the International Astronomical Union (of which he was President of the Commission on Celestial Mechanics from 1948 to 1955), and a foreign associate of the Royal Astronomical Society (England).

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R. T. Craig

Robert T. Craig was born on October 10, 1927, in Niagara Falls, N. Y. He received the B.S.E.E. degree from Massachusetts Institute of Technology, Cambridge, and the M.S.E.E. degree from Case Institute of Technology, Cleveland, Ohio. As a project engineer with the National Advisory Committee for Aeronautics from 1949 to 1955, he worked on the design and installation of controls and instrumentation for propulsion test facilities. He conducted research on control systems for gas turbine engines and wrote several NACA reports on gas turbine engine dynamic characteristics, control systems, and computer techniques.

He joined Thompson Ramo Wooldridge Inc., Cleveland, Ohio, in 1955 and directed design and development in advanced propulsion control systems and large-scale simulation studies. He has conducted system optimization and control analyses of nuclear power system and pneumatic controls, and has directed design of new analog computer components. He has directed trajectory and propulsion system analysis of electrically propelled vehicles, and has conducted research in high-temperature materials in the field of ion propulsion. He is now in charge of projects in propulsion system optimization, and experimental ion and plasma physics programs at the New Devices Laboratories, Thompson Ramo Wooldridge, Inc.

Mr. Craig is a member of AIEE, ARS, and Sigma Xi.

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W. R. Crone

W. Reed Crone (S'52–M'55) was born in Albert City, Iowa, on January 25, 1931. He attended the Milwaukee School of Engineering, Milwaukee, Wis., from 1948 to 1950, after which he served in the U. S. Navy as an electronic technician. In 1952, he was discharged to participate in the NROTC program at the University of Wisconsin, Madison, and received the B.S.E.E. and Bachelor of Naval Science degrees from that institution in 1955. He then re-entered the naval service and was assigned as Electrical Officer on an aircraft carrier until his release in 1958. Since that time, he has been pursuing graduate studies in electrical engineering and acting as research assistant at The Ohio State University Radio Observatory, Columbus, Ohio.

Mr. Crone is a member of Eta Kappa Nu and Tau Beta Pi and an associate member of the AIEE.

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A. F. Daniel

Arthur F. Daniel was born in Birmingham, Ala., on September 27, 1905. He received the B.S. degree in chemistry in 1927 and the M.S. degree in physics in 1928, both from the University of North Carolina, Chapel Hill. From 1928 to 1929 he attended the University of Chicago, Chicago, Ill., on a Charles A. Coffin fellowship in physics.

From 1929 to 1932, he was a research associate in electrical engineering at Harvard University, Cambridge, Mass. From 1933 to 1935, he was a research physicist with the General Chemical Co., Long Island, N. Y., and from 1935–1941, he was also a research chemist with the Storage Battery Division of T. A. Edison, Inc., West Orange, N. J. From 1941–1956, he was chemical engineer and Chief of the Battery Development Section of the U. S. Army Signal Research and Development Laboratory, Fort Monmouth, N. J. Presently he is a USASRDL general engineer and director of the Power Sources Division, Electronic Components Research Department, Fort Monmouth, where he is responsible for direction and coordination of research and design and development program to provide energy sources and energy conversion devices for use with U. S. Army Signal Corps equipment. In the field of primary batteries and special-purpose batteries, he is responsible for directing a research and development program for the Army and Air Force, and has received an Army exceptional civilian service award for battery research and development.

Mr. Daniel is a member of the American Chemical Society, the Electrochemical Society, the American Physical Society, the Armed Forces Communication Association, the Armed Forces Management Association, Sigma Xi, and Sigma Phi Sigma. He is listed in "American Men of Science."

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J. B. Day

John B. Day was born on March 3, 1928 in Erie, N. D. He served with the U. S. Navy as an airborne electronics technician from 1946 to 1951. In 1954 he received the B.A. degree in mathematics from San Jose State College, San Jose, Calif., and has studied graduate mathematics there and at Purdue University, Lafayette, Ind. After three years spent in teaching secondary mathematics in California and Indiana, he joined the Philco Western Development Laboratories, Palo Alto, Calif., in June, 1959. He is presently an engineer in the advanced applications section there.

Mr. Day is a member of Tau Delta Phi and Kappa Delta Pi.

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R. H. Dimond

Robert H. Dimond was born in Madison, Wis., on March 6, 1931. He received the B.S.E.E. degree in 1955 from the Milwaukee School of Engineering, Milwaukee, Wis.

He joined Radiation, Inc., Orlando, Fla., in 1956 as an engineer in the Instrumentation Division. His work there has included contributions in atomic effects and complete environmental testing of airborne electronic systems; FM/FM, PPM/AM, PCM/FM telemetry systems; oscillographs; tape recorders; and associated instrumentation. He is currently with the Research Division of Radiation, Inc., which he joined in 1958 as a senior staff engineer. Included in his work there have been military technical intelligence evaluations and systems engineering in radio telemetry.

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E. F. Dirsa

Edward F. Dirsa (M'56) was born in Brighton, Mass. on October 20, 1925. He received the B.S. degree in physics from Boston College, Chestnut Hill, Mass. in 1950, and the M.S. degree in physics from Indiana University, Bloomington, in 1952.

From 1952 to 1954, he was engaged in magnetron tube development with Sylvania Electric Products, Inc., Boston, Mass. From 1954 to 1958, he was associated with Cook Research Laboratories, Morton Grove, Ill, where he worked on studies of electromagnetic propagation, design and analysis of countermeasures equipment and radar systems, and design of a unique radar target range for the Air Force. He is presently engaged in military systems engineering with the Admiral Corporation, Chicago, Ill.

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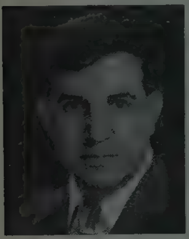

R. L. Easton

Roger L. Easton (M'46) was born in Craftsbury, Vt., on April 30, 1921, and received the B.A. degree in physics from Middlebury College, Vt., in 1943.

Since that time he has been at the U. S. Naval Research Laboratory, Washington, D. C. He has worked on radar beacons, IFF, and on rocket and satellite tracking equipment. He worked on the design of the Minitrack system and had design responsibility for the 1958 Beta 2 (Vanguard I) satellite. He is now responsible for the detection portion of the Navy's dark satellite tracking program as head of the Space Surveillance Branch, NRL.

Mr. Easton is a member of the ARS.

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D. H. Ellis

Donald H. Ellis (M'59) was born in Spartanburg, S. C., on March 8, 1927. He received the B.E.E. degree from Clemson College, Clemson, S. C., in 1951.

From 1944 to 1948 he served with the U. S. Navy. After graduating from college, he worked from 1951 to 1954 with the Atomic Energy Division of the E.I. du Pont deNemours Company on the control instrumentation of nuclear reactors. He then spent four years at the Massachusetts Institute of Technology's Lincoln Laboratory, Lexington, Mass., on the design and development of magnetic-core memory for digital computers. His prime responsibility was the development of a 256×256 coincident current magnetic-core memory for use in a large-scale digital computer, but he was also associated with the development of a high-speed transistorized core memory for use in a solid-state digital computer. At present, he is the Systems Development Unit Head involved in the design and development of airborne PCM data acquisition systems at Radiation, Inc., Melbourne, Fla.

Mr. Ellis is a member of Tau Beta Pi.

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C. W. Fetheroff

Charles W. Fetheroff was born on April 9, 1930, in East Cleveland, Ohio. He received the B.S.E.E. degree from Case Institute of Technology, Cleveland, Ohio, and has taken advanced work in servomechanisms, mathematics, circuit analysis, electronic circuits, electromagnetic theory, and atomic physics. He served in the U. S. Army Signal Corps from 1953 to 1955, doing research and development work on integrated radar defense systems.

Since joining Thompson Ramo Wooldridge, Cleveland, Ohio, in 1952, he has been responsible for electronic and magnetic component design, system stability studies for parallel alternators, design and development of auxiliary power unit simulators, and parametric analysis and startup stability studies on the Tartar and Terrier APU speed controls. He has performed hot gas servo analyses and has done research and development on electronic and mechanical components for application in extremely high ambient temperatures. In the field of advanced space propulsion, he has performed analysis and comparison of capabilities of chemical and electric propulsion systems, and is now in charge of refined studies on performance comparison of these two systems.

Mr. Fetheroff is a member of AIEE, ARS, Tau Beta Pi, Sigma Xi, and Eta Kappa Nu.

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J. J. Fleming

James J. Fleming (A'42–M'55) was born in Chicago, Ill., on February 26, 1917. He received the B.S. degree in physics at Northwestern University, Evanston, Ill., in 1938. In 1941, after three years of graduate study in physics and service as a research assistant at Northwestern, he joined the Radio Division of the U. S. Naval Research Laboratory, Washington, D. C. During World War II, he contributed to the development of plan position indication for search radars. He was also responsible for the operational evaluation of Navy anti-aircraft gun fire-control systems, for which he received the Meritorious Civilian Service Award.

Since 1945, he has been Head of the Operational Research Branch at the U. S. Naval Research Laboratory, first in Radio Division III and now in the Applications Research Division. In addition to the development of the NRL Electronic Digital Computer (NAREC), the program of the Branch has included, for Project Vanguard, the development of an automatic telemetered data reduction facility (ARRF), a radar-computer ground-controlled third-stage firing system, and the Vanguard Computing Center. Currently, the Branch is engaged in the data processing problem of the Space Surveillance program.

Mr. Fleming is a member of the American Physical Society, the Association for Computing Machinery, the Scientific Research Society of America, and Phi Beta Kappa.

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R. Franklin

Rachel Franklin was born in Dudley, N. J., on March 27, 1897. She received the B.A. degree from Vassar College, Poughkeepsie, N. Y., in 1919, and the M.A. degree from the University of Pennsylvania, Philadelphia, in 1929.

Her main interests have been in the field of optics and in the design of optical instruments. From 1930 to 1942, she was associated with the Biochemical Research Foundation in Newark, Del., working on the effects of ultraviolet radiation on living tissues and protozoa, and on the absorption spectra of normal and cancerous tissues. During World War II, she was employed by the U. S. Army Signal Corps, Ft. Monmouth, N. J., as an engineer in the development and inspection branch. In 1946, she joined the staff of the Franklin Institute Laboratories, Philadelphia, Pa., as a research physicist. She has worked in that capacity on various projects involving optics, including ultraviolet and infrared absorption spectroscopic studies and the design of specialized instruments for this work.

Miss Franklin is a member of the Optical Society of America, Phi Beta Kappa, and the Scientific Research Society of America, Franklin Institute Branch.

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W. H. Guier

William H. Guier was born on July 13, 1926, in Wichita, Kans. He received the B.S. degree in 1948, the M.S. degree in 1950 and the Ph.D. degree in theoretical physics in 1951, all from Northwestern University, Evanston, Ill.

He joined the Applied Physics Laboratory of The Johns Hopkins University, Silver Spring, Md., in 1951, and is now a principal staff member in the field of theoretical physics. He recently received the "Outstanding Young Scientist" award which is sponsored by the Maryland Academy of Science.

Dr. Guier is a member of Lambda Chi Alpha, the American Physical Society, and the Philosophical Society of Washington, D. C.

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William R. Haseltine was born in Chicago, Ill., on June 27, 1911. He received the B.S. and Ph.D. degrees in physics in 1934 and 1938, respectively, from the Massachusetts Institute of Technology, Cambridge, Mass. He was a teaching fellow at M.I.T. in 1935 and 1936, an assistant in mathematics at the University of Wisconsin, Madison, in 1936 and 1937, and a research fellow in physics and lecturer at the University of California, Berkeley, from 1938 to 1941. From 1941 to 1945, he was an officer in the U. S. Army Ordnance Department.

W. R. HASELTINE

Since 1946, he has been a physicist, primarily in rocket ballistics, at the U. S. Naval Ordnance Test Station, China Lake, Calif.

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Robert C. Higgy (A'26–SM'45) was born in Columbus, Ohio, on December 7, 1901. He received the B.S.E.E. degree from The Ohio State University, Columbus, in 1925 and the Professional Electrical Engineering degree in 1937.

R. C. HIGGY

Since 1925, he has been employed by The Ohio State University as Director of Broadcasting Stations, Associate Professor of Electrical Engineering, and since 1956 as Associate Director for Engineering for radio and television stations and Assistant Director of the Radio Observatory. Recently, his work has included teaching electrical engineering and research work related to ionization effects due to earth satellites. He is the author of a book on fundamental radio experiments and of articles on ground wave propagation.

Mr. Higgy is a member of Sigma Xi and a licensed Professional Engineer in Ohio.

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Donald C. Hock was born in North Tonawanda, N. Y., on April 18, 1929. He received the B.A. degree from the University of Buffalo, N. Y., in 1951, the M.S. degree from the A & M College of Texas, College Station, in 1952, and the Ph.D. degree in physics from Pennsylvania State University, University Park, in 1956.

D. C. HOCK

He joined the Applied Physics Laboratory of The Johns Hopkins University, Baltimore, Md., in 1955 as a member of the Senior Staff, and since 1958, has been with the Research Division of Radiation Incorporated, Orlando, Fla.

Dr. Hock is a member of Sigma Pi Sigma, Phi Kappa Phi, Pi Mu Epsilon, and Sigma Xi.

Conrad H. Hoeppner (SM'47–F'58) was born on March 12, 1918, in Spooner, Wis. He received the B.S.E.E. degree in 1939 and the M.S.E.E. degree in 1940, both from the University of Wisconsin, Madison. He did graduate work at the Massachusetts Institute of Technology, Cambridge, and received the professional E.E. degree in 1947 from the University of Wisconsin.

C. H. HOEPPNER

Thereafter, he was employed by the U. S. Naval Research Laboratories, Washington, D. C.; he was Director of the Electronics Laboratory, Glenn L. Martin, Co., Baltimore, Md.; Director of the Engineering Products Department, Raytheon Manufacturing Company, Waltham, Mass.; Vice President and Director of General Electronics Laboratory, Inc., Cambridge, Mass.; Manager of the Electronic Division, W. L. Maxson Corp., New York, N. Y.; and Manager of the Development Division, Stavid Engineering, Plainfield, N. J. He is presently Chief Scientist at Radiation, Inc., Melbourne, Fla.

Mr. Hoeppner is a member of Sigma Xi, Pi Mu Epsilon, Phi Kappa Phi, Phi Eta Sigma, Tau Beta Pi, and Eta Kappa Nu, and is a Fellow of Tau Beta Pi, and the Wisconsin Alumni Research Institute.

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Hyman Hurvitz (M'53) was born in Boston, Mass., on February 18, 1910. He received the B.E.E. degree from Northeastern University, Boston, in 1930, and the L.L.B. and M.P.L. (Master of Patent Law) degrees from Washington College of Law (now American University), Washington, D. C., in 1936 and 1937, respectively.

H. HURVITZ

He became a member of the Washington, D. C Bar in 1937, and from 1942 to 1945, he served in the U. S. Navy, rising to the position of Lieutenant Commander. Since that time he has been in private practice, presently with the firm of Hurvitz and Rose, Washington, D. C.

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H. P. Hutchinson (A'32–SM'46) was born in Brooklyn, N. Y., in 1908. He received the B.S.E.E. degree from the Polytechnic Institute of Brooklyn, Brooklyn, N. Y., in 1930 and the M.S.E.E. degree from Columbia University, New York, N. Y., in 1933.

He has held numerous scientific and engineering positions, the most recent being Research Engineering, National Bureau of Standards; Chief Engineering, U. S. Army Aviation Center; and Assistant Director of Research, U. S. Army Signal Research and Development Laboratory, Ft. Monmouth, N. J. During World War II and in the Korean emergency, he served on military duty at the Pentagon directing Signal Corps' participation in nuclear weapons tests and in applied research.

H. P. HUTCHINSON

Commencing with the first Sputnik, he has been continuously engaged in research on satellite signals and, together with P. R. Arendt, has written several papers on satellite radio propagation, His present position is Special Assistant to the Chief, Applied Physics Division, U. S. Army Signal Research and Development Laboratory.

Mr. Hutchinson is an Associate Member of Commission 3, USA National Committee of URSI.

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Donald J. Jacoby (M'46) was born in New York, N. Y., on July 8, 1920. He received the B.S.E.E. degree from Columbia University, New York, in 1941, and the M.S.E.E. degree from Rutgers University, New Brunswick, N. J., in 1950.

D. L. JACOBY

Since 1941, he has been employed at the U. S. Army Signal Research and Development Laboratory, Fort Monmouth, N. J., with the exception of three years as a radar officer in the U. S. Navy. He has been engaged in research and development programs in fields of multiplexing, radio relay and tropospheric scatter, and until recently was chief of the Radio Relay Branch of the Transmission Facilities Division. He is currently deputy director of the Astro-Electronics Division, USASRDL, with responsibility for directing programs in satellite communications, satellite instrumentation and tracking. He holds a patent on "Pulse Code Modulation."

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James R. Jenness, Jr. was born in Atlanta, Ga., on February 5, 1925. He received the B.S. degree in physics and mathematics from Parsons College, Fairfield, Iowa, in 1945 and the M.A. degree in physics from Temple University, Philadelphia, Pa., in 1954.

J. R. JENNESS, JR.

In 1945–1946, he taught mathematics in the high school at Birmingham, Iowa, was appointed to a graduate assistantship in physics at the

State University of Iowa, Iowa City, in 1946–1947, and in 1947–1948, was a senior fellow in physics at Iowa State College, Ames. From 1948 to 1951, he worked on special instrument calibration problems and the development of aeronautical research instruments for the NACA at Langley Field, Va. From 1951 to 1956, he was a physicist at the U. S. Naval Air Development Center, Johnsville, Pa., where his field of interest was infrared optics and guidance systems. Since 1956 he has been a Senior Physicist at HRB-Singer, Inc., State College, Pa., where he introduced the Hass-Erbe-Bradford replica mirror process for adaptation to large-scale production of low-cost precision reflective optical components by the Singer Military Products Division, and also developed techniques for fabricating paraboloidal microwave antennas by centrifugal casting. He has had a continuing interest in research instrumentation and the functioning of instruments and electronic components under unusual environmental conditions.

Mr. Jenness is a member of Phi Kappa Phi, Sigma Pi Sigma, Chi Beta Phi, the Optical Society of America, and the Association for Applied Solar Energy.

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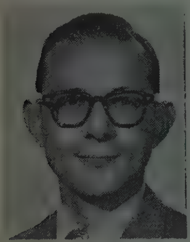
Conrad S. Josias (S'50,–A'51,–M'56) was born in New York, N. Y., on June 12, 1930. He received the B.E.E. degree from New York University, New York, in 1951 and the M.E.E. degree from the Polytechnic Institute of Brooklyn, N. Y., in 1955.


C. S. Josias

He was employed by Airborne Instruments Laboratory, Mineola, N. Y., from 1951 to 1956, during which time he was active in the development of electronic devices and systems. In 1956, he joined the Jet Propulsion Laboratory, California Institute of Technology, Pasadena, where he participated in the development of transistor circuits for missile guidance systems and space payload instrumentation. At present, he is a senior research engineer in the JPL Space Instruments Section.

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Sidney W. Kash was born on March 1, 1922, in New York City, N. Y. He received the Ph.D. degree in physics in 1951, from the University of California at Los Angeles.

During World War II, he worked at the Naval Ordnance Laboratory in Washington, D. C., on underwater degaussing problems. In 1947, he participated in a spectroscopic program related to the V2 firings at White Sands, N. M. In the early part of 1951, he designed equipment for the automatic measurement of skylight polarization for the Meteorology Department at UCLA. For the next four years, he was a senior engineer at Atomics International in Canoga Park, Calif., engaged in research and development of thermal neutron reactors. He has been with Lockheed Aircraft Corporation, Palo Alto., Calif., since the latter part of 1955, and is presently manager of the Ionic Physics Department. While at Lockheed, he has been engaged in a number of activities, including the design of high-pressure and electromagnetic shock tubes and research in plasma physics, hydromagnetics, and electrical propulsion.

S. W. Kash

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P. E. Kendall (SM'56) was born in Paoli, Ind., on July 27, 1921. He received the B.S.E.E., M.S.E.E. and Ph.D. degrees from Purdue University, Lafayette, Ind., in 1948, 1949 and 1953, respectively.

P. E. Kendall

He has taught undergraduate courses in electrical engineering at Purdue, and is an affiliate assistant professor at that university, teaching graduate courses in automatic controls and pulse circuitry. At ITT Laboratories, Fort Wayne, Ind., he has worked on system analysis and design of missile and space vehicle guidance and navigation systems; research in the field of nonlinear mechanics as applied to servomechanisms; optimization of contactor servos, and mercury drop noise generators in microwave frequencies; analysis of space vehicle trajectories; development of navigational equipments including Loran and radar altimeters; and operation and maintenance of military communications and radar equipments. He has held administrative positions as Head of the Electromechanical Section of the Advanced Development Department, Head of the Systems Analysis and Design Section of the Astrionics Laboratory, the position which he now holds.

Dr. Kendall is a member of Eta Kappa Nu and Sigma Xi. He has two patents pending.

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Robert L. Kent (M'59) was born in Philadelphia, Pa., on May 25, 1929. He attended the Moore School of Electrical Engineering, University of Pennsylvania, Philadelphia, where he received the B.S. degree in 1950, and the Massachusetts Institute of Technology, Cambridge, Mass., where he received the M.S. degree in 1952.

From 1950 to 1952, he was a research assistant in electrical engineering at M.I.T. Since that time, he has been a staff member of the Research Laboratory of Electronics, M.I.T., participating in the development of missile guidance and electronically scanned radar systems.

R. L. Kent

Mr. Kent is a member of Eta Kappa Nu, Sigma Tau, and Tau Beta Pi, and an associate of Sigma Xi.

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John D. Kraus (A'32–M'43–SM'43–F'54) was born in Ann Arbor, Mich., on June 28, 1910. He received the B.S. degree in 1930, the M.S. degree in 1931, and the Ph.D. degree in 1933, all from the University of Michigan, Ann Arbor.

J. D. Kraus

From 1934 to 1935, he did research on industrial noise-reduction problems and from 1936 to 1937, was engaged in nuclear research with the newly completed University of Michigan cyclotron. From 1938 to 1940, he was an antenna consultant. In 1940, he joined the Naval Ordnance Laboratory, Washington, D. C., working on the degaussing of ships, and in 1943, he became a member of the Radio Research Laboratory at Harvard University, Cambridge, Mass. In 1946, he joined the faculty of The Ohio State University, Columbus, where he is now professor of electrical engineering, professor of physics and astronomy, and Director of the Radio Observatory. He is responsible for the development of the helical beam antenna, the corner-reflector antenna, and other antenna types.

Dr. Kraus is a member of the American Astronomical Society and the American Physical Society.

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Leonard R. Malling (A'31–VA'39–M'59) was born in London, Eng., on July 9, 1909. He graduated from Northampton Technical Institute, London, Eng., in 1931 and came to this country in 1938.

L. R. Malling

For some 30 years, he has done research and design engineering in such areas as radar, nuclear resonance, aircraft guidance, guided missiles, and space vehicles. During this time, he has been employed by Hazeltine Electronics Corporation, Boeing Airplane Company, Convair, Varian Associates, and the U. S. Naval Laboratory, San Diego, Calif. In 1955 he joined the staff of the Jet Propulsion Laboratory, California Institute of Technology, Pasadena, where he is presently a senior research engineer.

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F. N. McMillan

F. N. McMillan was born on May 26, 1923, in Fayetteville, N. C. He received the B.S. degree with a group major in chemistry, mathematics and German from the University of Florida, Gainesville, in 1947 and the M.S. degree in chemistry from the same school in 1949.

From 1950 to 1955, he served as a chemist in the Florida Food and Drug Laboratory in Tallahassee. While in Tallahassee, he did additional graduate work in chemistry at Florida State University. From 1955 to 1958, he served as a research chemist at Eglin Air Force Base, Fla. Since 1958, he has been with the Research Division of Radiation Incorporated, Orlando, Fla. His work has included chemical kinetics, spectroscopy, immunology, organic synthesis, incendiaries, instrumentation, re-entry, and rocket propulsion.

Mr. McMillan is a member of Sigma Xi and the American Chemical Society.

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H. F. Meyer

Harold F. Meyer (SM'50) was born in New York, N. Y., on September 2, 1916. He attended Boston University, Boston, Mass., and Columbia University, New York, N. Y.

For many years he was active as a radio experimenter, amateur, member of the Naval Communication Reserve, and in the radio business. Entering Government service in the Signal Corps Plant Engineering agency in 1942, he became Chairman and Technical Secretary of the Technical Coordination Committee and Chief of the Engineering Coordination Section. In 1945, he transferred to the U. S. Army Research and Development Laboratories, Ft. Monmouth, N. J., and became Chief Engineer of the Long-Range Equipment Section. In 1955, he was made Chief of the Long-Range Radio Branch. In these capacities from 1946 to 1957, he was responsible for research and development of the Army long-distance radio communication equipment and systems. He joined the RamoWooldridge Division of Thompson Ramo Wooldridge Incorporated, Los Angeles, Calif., in 1957 as Co-Head of the Communications Systems Department and is currently employed in that capacity.

A. B. Moody

Captain Alton B Moody, USNR, was born in Thatcher, Ariz., on July 28, 1911. He graduated from the U. S. Naval Academy, Annapolis, Md., in 1935.

He has had experience in marine navigation in both the Navy and merchant marine, and in air navigation in both the Navy and Air Force. He has taught navigation in private schools and at the Naval Academy. He has been at the U. S. Navy Hydrographic Office since 1946, and for several years has been Deputy Director of the Navigational Science Division. In 1953 he received the international Thurlow Navigation Award for the outstanding contribution to the science of navigation.

Captain Moody is now President of the Institute of Navigation, a government member of the Governing Board of the U. S. Power Squadrons, a member of the Board of Governors of the American Polar Society, and a member of the New York Academy of Sciences and the Arctic Institute of North America.

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W. L. Morrison

William L. Morrison (A'52–M'59) was born in Salt Lake City, Utah, on April 3, 1924. After three years as an electronics technician in the U. S. Navy, he attended the University of California, Berkeley, and received the B.S.E.E. degree in 1953.

From 1953 to 1955 he was an electronic scientist at the U. S. Naval Radiological Defense Laboratory, San Francisco, Calif. From 1956 to 1958 he assisted in the design of the first nuclear reactor control console for the Atomic Energy Division of American-Standard, Inc., Mountain View, Calif. He is presently a senior research engineer in the Missiles and Space Division of Lockheed Aircraft Corporation, Palo Alto, Calif.

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G. E. Mueller

George E. Mueller (S'39–A'41–SM'46) was born on July 16, 1918, in St. Louis, Mo. He received the B.S.E.E. degree from the Missouri School of Mines, Rolla, in 1939. He continued his education at Purdue University, Lafayette, Ind., where he was a Research Fellow, and obtained the M.S.E.E. degree in 1940. He studied physics at Princeton University, Princeton, N. J., and received the Ph.D. degree in physics from The Ohio State University, Columbus, in 1951. He served as professor of electrical engineering at The Ohio State University for more than ten years.

His technical experience has included television and microwave research at Bell Telephone Laboratories, Holmdel, N. J., and work as a consultant for the Ramo-Wooldridge Corporation, Los Angeles, Calif. He is the author of nine technical publications, and holds six patents in the electron tube and antenna fields. He is presently Vice-President and Associate Director of the research and development division, Space Technology Laboratories, Inc., Los Angeles, Calif.

Dr. Mueller is a member of AIEE and the American Physical Society.

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H. L. Newkirk

Horace L. Newkirk was born in Minneapolis, Minn., on June 11, 1911. He received the B.S. degree from Union College, Schenectady, N. Y., in 1933 and the M.S. degree from the University of Idaho, Moscow, in 1940, both in physics. He has since taken additional courses in mathematical physics through the extension program of the University of California at Los Angeles.

From 1940 to 1945 he worked as a physicist in the Navy degaussing program. Since 1945, he has continued as a physicist with the Navy at the Naval Ordnance Test Station, China Lake, Calif., primarily in aeroballistics of rockets.

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G. C. Newton, Jr.

George C. Newton, Jr. (M'47) was born in Milwaukee, Wis., on May 14, 1919. He received the B.S.E.E. degree in 1941 and the D.Sc. degree in electrical engineering in 1950, both from the Massachusetts Institute of Technology, Cambridge. During World War II, he was engaged in defense research and development work at the Special Ordnance Plant in York, Pa., and at the Sperry Gyroscope Company, Inc., Long Island, N. Y.

He is presently Associate Professor of electrical engineering and Associate Director of the Electronic Systems Laboratory at M.I.T. In addition to teaching fundamental electrical engineering subjects and advanced courses in automatic control, he has research interests in the areas of instrumentation, control and computation. His consulting activities include services to both government and industry on military and commercial problems involving electronics and control technology.

Dr. Newton is a member of the AIEE, the ASEE, and the Franklin Institute. He received Franklin Institute Levy Medal in 1953.

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R. R. Newton

Robert R. Newton was born on July 7, 1918 in Chattanooga, Tenn. He attended the University of Tennessee, Knoxville, where he received the B.S.E.E. degree in 1940. He received the Ph.D. degree in theoretical physics from The Ohio State University, Columbus, in 1946.

For several years he taught in the Physics Departments of the University of Tennessee and Tulane University, New Orleans, La. During this time he also directed a research program in theoretical exterior ballistics for the Ordnance Missile Laboratories, Redstone Arsenal. He joined the Applied Physics Laboratory of The Johns Hopkins University, Silver Spring, Md., in August, 1957, and is now a principal staff member in the field of theoretical physics and supervisor of the Space Exploration Group.

Dr. Newton is a member of the American Physical Society, ARS, Sigma Xi, and Tau Beta Pi.

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W. A. Nichols

William A. Nichols (M'57) was born in Spartanburg, S. C., on December 28, 1921. He served in the U. S. Marine Corps from January, 1940 to June, 1948 as radio-radar technician and instructor. He was an instructor in the U. S. Navy Electronics Technician Training Program from 1944 to 1947.

He joined the U. S. Naval Research Laboratory, Washington, D. C., in November 1948. He has been responsible for the design and development of electronic instruments and telemetering facilities to implement rocket and satellite probes studying radiations of solar and other celestial bodies in the ultraviolet and X-ray regions. At present, he is Head of the Electrical Design and Engineering Section of the Upper Air Physics Branch.

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Joseph C. Nowell was born in Boston, Mass., on December 18, 1924. He received the B.S. degree in 1948, and the M.S. degree in 1950, both from Massachusetts Institute of Technology, Cambridge, Mass. He remained at M.I.T. to work on missile control systems in the Dynamic Analysis and Control Laboratory, and subsequently on missile guidance and advanced radar techniques in the Research Laboratory of Electronics. He is now with the Advanced Development Department of the Missile Systems Division, Raytheon Manufacturing Company, Bedford, Mass.

J. C. Nowell

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D. D. Ordahl

Douglas D. Ordahl was born on April 17, 1921, in Santa Rosa, Calif. He received the B.S. degree in chemistry from Stanford University, Stanford, Calif., in 1942, and the M.S. degree in chemical engineering from The Ohio State University, Columbus, in 1948.

He entered the propulsion field in 1942 at the Hercules Powder Company, Radford, Va., and in 1945 became a research engineer at Battelle Memorial Institute, Columbus, Ohio. In 1950, he joined the Naval Ordnance Test Station, China Lake, Calif., in order to continue research on the physical properties of solid propellants. He is presently Associate Head of the Propulsion Development Department there and does research, design, development and program direction of solid-, liquid-, ramjet-, and hybrid-propulsion systems.

Mr. Ordahl is a member of Phi Lambda Upsilon, ARS, and the American Chemical Society.

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M. B. Pickover

Merwin B. Pickover was born in New York, N. Y., October 8, 1929. He received the B.E.E. degree from the College of the City of New York in 1951.

He was project engineer with the Bendix Corporation, Teterboro, N. J., and was transferred to the Red Bank division in 1952. His responsibility as project engineer was the design of missle and aircraft electrical accessories. In 1958, he joined the U. S. Army Signal Research and Development Laboratory, Fort Monmouth, N. J. As senior engineer, he now heads the team engaged in the design and development of tape recorder circuitry for satellite instrumentation; particularly, transistorized controls on recorder motors, for specialized applications as well as associating circuitry. At this time he is assigned to the Astro-Instrumentation Branch, Astro-Electronics Division, USASDRL.

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N. S. Potter

Norman S. Potter (SM'58) was born in New York, N. Y., on November 5, 1926. He received his undergraduate training at Brooklyn College, Brooklyn, N. Y., and at the U. S. Naval Academy, Annapolis, Md., where he studied marine and electrical engineering. He received the B.A. degree in mathematical physics from Brooklyn College in 1945 and the M.S. degree in applied mathematics and physics from Massachusetts Institute of Technology, Cambridge, in 1953. He continued supplementary studies at M.I.T. and Columbia University, New York, N. Y., for the Doctorate Degree.

As a member of the research staff, Columbia University, he was associated with the Manhattan Project, where he performed analytic studies in radiation propagation. Later a member of the research staff of the Department of Electrical Engineering at M.I.T., he worked for some years on SAGE system development, electron digital computer logic, data handling in real-time problems, advanced target acquisition techniques, and general air defense systems analyses. Following this, he was affiliated with The Martin Company for several years as a senior engineer in the field of advanced design, with primary responsibility for systems studies in several categories of missiles, airborne fire control systems, missile guidance techniques, and the analysis of composite airborne electronics subsystems.

He then served as a consultant at USAF Headquarters in Europe in the field of technical evaluation of foreign airborne weapons systems. Following his return from Europe, he was with the Air Arm Division of Westinghouse where, for several years as a Fellow Engineer and Weapons Systems Group Leader, he had primary responsibility for systems analyses and development in air-launched guided missiles, bomber detection and penetration techniques, airborne early warning and interceptor-borne detection and tracking equipments, advanced automatic acquisition techniques in discrete data systems, and operations analysis in large-scale air defense system evaluations.

For some years he has been engaged as a special consultant on guided missiles systems to the Assistant Chief of Staff, Intelligence, Department of the Army, and is presently manager of the Weapons Systems Research Laboratory of the Research and Development Division of the W. L. Maxson Corporation, New York, N. Y., where he is concerned with advanced systems studies in all phases of military and industrial electronics.

He is a member of the IAS, ARS, and AAAS.

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Julius Praglin (M'56) was born in Boston, Mass., on June 12, 1927. He received the B.A. degree in biology from Clark University, Worcester, Mass., in 1948, and the M.S. and Ph.D. degrees in physiology from the University of Illinois, Urbana, in 1950 and 1952, respectively.

J. Praglin

From 1952 to 1955, he was engaged in research on the biophysics of vision at Western Reserve University, Cleveland, Ohio. In 1955, he joined Keithley Instruments, Inc., Cleveland, Ohio, as director of Research and Development. He became a Vice-President of Keithley Instruments in 1957 and presently is in charge of Research and Engineering.

Dr. Praglin is a member of Sigma Xi.

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A. V. Pratt was born in Maupin, Ore., on July 4, 1924. He received the B.A. degree from the University of California, Berkeley, in 1950 and has since done graduate work at the University of Utah, Salt Lake City, and the University of Colorado, Boulder.

A. V. Pratt

He has worked for the U. S. Naval Ordnance Test Station, China Lake, Calif., since 1950 and is presently serving as a physicist in the Weapons Development Department there.

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Alfred G. Ratz (M'49) was born in Hamilton, Can., on April 20, 1922. He received the B.A.Sc. degree in 1944, the M.A.Sc. degree in 1947, and the Ph.D. degree in 1951 from the University of Toronto, Can.

A. G. Ratz

From 1947 to 1949, he was lecturer in electrical engineering at the University of Toronto. For the next three years, he was project engineer at the University's Computation Center where he did basic development work on digital computer techniques, culminating in the building of a digital computer. From 1952 to 1957, he was engaged in airborne electronic systems work at the Air Armament Engineering Department of Canadian Westinghouse. As a project engineer on air-to-air missile electronics, he was responsible for the development circuitry, subminiaturization, ruggedization, reliability and packaging of guidance systems. Subsequently, as section engineer and manager of the Systems Section, he directed a wide range of airborne projects, including guidance, control, aerodynamics, countermeasures, and planning and execution of missile flight tests and firings. In 1957, he joined the Applied Science Corporation of Princeton, Princeton Junction, N. J., where as assistant manager of the Engineer Planning Department he directed the study, research and development of pioneer telemetering hardware, including statistical telemetering devices. In January, 1959, he was appointed chief engineer. He joined Gulton Industries, Inc., in June, 1959, as Chief Engineer of the new Ortholog Division specializing in instrumentation, telemetry, and data-handling systems.

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Henry B. Riblet (SM'44) was born on May 20, 1911, in Clayton, N. M. He graduated from Friends University, Wichita, Kans. in 1934 with a B.A. degree in physics and chemistry.

H. B. Riblet

After graduation, he worked at radio station KLZ in Denver, Colo., as chief transmitter engineer. In 1940, he joined the consulting radio engineering firm of Glenn D. Gillett, Washington, D. C., where he was engaged in design and adjustment of directional antennas for radio broadcasting stations. During World War II, he was a member of the Research and Development Staff at Airborne Instruments Laboratory directed by Columbia University, New York, N. Y., where he was engaged in the development of airborne magnetic detection for submarines. After the war, he rejoined the Glenn D. Gillett firm. In March, 1949, he joined the staff of The Johns Hopkins University, Applied Physics Laboratory, Silver Spring, Md., to work with the Telemetering Group. In 1950, he was appointed Group Supervisor of the Telemetering Group, later known as Instrumentation Development. Recently, he has been the supervisor of an Electronics Group conducting development and engineering for satellite instrumentation.

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Robert W. Rochelle (S'46–A'50–M'58) was born on June 23, 1923, in Nashville, Tenn. He received the B.S.E.E. degree from the University of Tennessee, Knoxville, in June, 1947, and the Master of Engineering degree from Yale University, New Haven, Conn., in 1949. He began graduate studies at the University of Maryland, College Park, and taught in their graduate program for two years.

R. W. Rochelle

In March, 1949, he joined the Naval Research Laboratory, Washington, D. C., where he participated in the development of electronic instrumentation for the Pacific nuclear test. In 1956, he joined the Magnetic Amplifier Section, where he designed a telemetry system for the Vanguard satellite program. In November, 1958, he was transferred to the National Aeronautics and Space Administration and became head of the Flight Data Systems Branch, Payload Systems Division. He is currently working on electronic instrumentation for satellites and space probes.

Mr. Rochelle is a member of Tau Beta Pi, RESA, and AAAS.

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Joseph L. Ryerson (A'55–M'55) was born on October 20, 1918, in Goshen, New York. He received the B.E.E. degree from Thomas S. Clarkson College of Technology, Potsdam, N. Y., in 1941 and the M.A.E.E. degree from Syracuse University, Syracuse, N. Y.

J. L. Ryerson

Upon graduation he was employed as a development engineer by the Ward Leonard Electric Co. of Mt. Vernon, N. Y. In 1946, he was employed by the Associated College of Upper New York as a Professor of physics and later transferred to Evansville College, Evansville, Ind., as a Professor of physics and electrical engineering. Here, in addition to teaching, he became a registered professional engineer and entered the consulting field. Employed as an electronics engineer by the Rome Air Development Center, Rome, N. Y., in 1951, he was assigned the problem of investigating automatic landing. He is now the Senior Scientist, Office of Advanced studies, there.

Mr. Ryerson is a member of Sigma Phi Sigma.

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Charles Saltzer (M'56) was born on February 13, 1918, in Cleveland, Ohio. He studied mathematics and physics at Western Reserve University, Cleveland, Ohio, the University of Illinois, Urbana, the University of Nebraska, Lincoln, and Brown University, Providence, R. I. At Brown University he received the doctorate degree in applied mathematics in 1949.

He worked in gas dynamics at Brown University from 1943 to 1945 and was an instructor of mathematics from 1945 to

1948. He was appointed instructor of mathematics at Case Institute of Technology, Cleveland, in 1948 and subsequently to the positions of assistant professor and associate professor. He is on partial leave of absence and holds the position of Consulting Engineer at Thompson Ramo Wooldridge Inc., Cleveland. He was appointed professor of applied mathematics at the University of Cincinnati, Ohio, in 1960.

C. SALTZER

He was retained by the General Electric Company as a consultant in information theory and computer design from 1956 to 1958. He has also done consulting work in numerical analysis, biophysics and space technology, and has published papers on electrical network theory, switching theory, distributions and numerical analysis. He spent a year in England (1950–1951) on a Fulbright Exchange. He received a National Science Foundation Award for research on Finite Difference Operators in 1959.

Dr. Saltzer is a member of the American Mathematical Society, the American Mathematical Association, the London Mathematical Society, the Association for Computing Machines, the Society for Symbolic Logic, the ARS, and Sigma Xi.

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W. F. SAMPSON

William F. Sampson was born in Oak Park, Ill., on July 6, 1929. He received the B.S. degree in physics from the California Institute of Technology, Pasadena, in 1951.

He then joined the research staff of Jet Propulsion Laboratories at the California Institute of Technology. During his six years there, he took a leading part in the development and field testing of the Microlock Receiving System used in the Explorer satellite program. In 1957, he became head of the Space Communications section of Hallamore Electronics Company, in Anaheim, Calif. He directed the development of several subsystems for advanced telemetry systems; supervised the design, installation and check-out of tracking stations; and directed systems integration projects for space probe tracking and guidance complexes. He was then promoted to engineering manager, and directed research programs and corporate projects related to advanced systems. In November, 1959, he joined Space Technology Laboratories, Los Angeles, Calif., where he is a senior staff engineer in the Telecommunications Laboratory.

Mr. Sampson is a member of Tau Beta Pi and of Sigma Xi.

A. A. SANDBERG

Allen A. Sandberg was born in New York, N. Y., on August 15, 1938. A senior at Massachusetts Institute of Technology, Cambridge, Mass., he expects to receive the B.S.E.E. and M.S. degrees in June, 1961.

Through participation in the electrical engineering, cooperative program at M.I.T., he was a student member of the AFCRC Astrosurveillance Sciences Laboratory in 1958 and of the AFCRC Plasma Physics Unit in the summer of 1959.

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R. W. SANDERS

Ray W. Sanders (SM'59) was born in Pomona, Calif., on April 24, 1927. He received the B.S. degree in physics and mathematics from Stanford University, Stanford, Calif., in 1948, and the M.S.E.E. degree from Stanford in 1950. From 1950 to 1951, he was an acting instructor and research assistant at Stanford.

From 1951 to 1958, he worked at Gilfillan Bros., Inc., Los Angeles, Calif. in the fields of missile systems analysis and design, missile flight tests, air traffic control systems, automatic electronic countermeasure systems, and air navigation system analysis. While at Gilfillan, he served as Director of the Missile Section in the Engineering Department. He is presently with Space Electronics Corporation, Glendale, Calif., where he is Manager of the Space and Satellite Systems Department.

Mr. Sanders is a member of Sigma Xi and the American Ordnance Association.

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C. L. SEARLE

Campbell L. Searle (S'46–A'48–M'55) was born in Winnipeg, Can., on July 24, 1926. He received the B.S.E.E. degree at Queens University, Kingston, Ontario, Can., in 1947, and the M.S. degree at the Massachusetts Institute of Technology, Cambridge, in 1951.

From 1948 to 1955, he was engaged in electronics research in the guided missile field at the Research Laboratory of Electronics, M.I.T., first as a research assistant and later, as a staff member of the Division of Sponsored Research. He joined the M.I.T. teaching staff in 1956, and is now an assistant professor in the Department of Electrical Engineering.

Mr. Searle is a member of Sigma Xi.

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G. F. SENN

George F. Senn (SM'50) was born on May 11, 1913 in Philadelphia, Pa. He received the B.S.E.E. degree from Drexel Institute of Technology, Philadelphia, in 1935.

In 1935 he was with Philco Radio and Television Co. in Philadelphia and in 1936 with Radio Corporation of America in Camden, N. J. In 1937, he joined the U. S. Army Signal Research and Development Laboratory at Fort Monmouth, N. J., and since that time has been Chief of the Vehicular Installation Section, Chief of the Quality Control Division, and is currently the Chief Engineer of the Communications Department. In 1958, he was the project manager in charge of the "SCORE" Communications Satellite Project.

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A. E. SHERBURNE

Arthur E. Sherburne (S'49–A'51–M'56) was born in Tyngsboro, Mass., on August 20, 1923. He received the B.S.E.E. degree from Tufts University, Medford, Mass., in 1950 and the M.S.E.E. degree from Massachusetts Institute of Technology, Cambridge, in 1951.

Since 1951, he has been employed at Trans-Sonics, Inc., Burlington, Mass., where he contributed to the development of airborne radar systems and in 1956 supervised the flight testing of a radar system. In 1958, he directed the development of an acoustic phase detector for Lox and JP-4. He has been a member of the New Product Research Department since its creation in 1958.

Mr. Sherburne is a member of Tau Beta Pi and Sigma Xi.

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Frederick F. Slack was born in Lawrence, Mass., on June 16, 1917. From 1941 to 1945, he was engaged in research and development at M.I.T. Radiation Laboratory, Cambridge, Mass., and was granted several patents on radar indicator circuits. In 1945, he joined a newly formed organization which is now known as the Air Force Cambridge Research Center. As section chief in the Air Traffic Control Laboratory, he was responsi-

F. F. Slack

ble for the design and development of the first complete automatic track-while-scan unit for the VOLSCAN system. In connection with this work he invented the "light gun" for the acquisition of radar targets and assignment of automatic tracking gates. The light gun has since been adopted by a number of agencies for application in military radar systems, among them the SAGE system.

He is now chief of the Development Section of the AFCRC Astrosurveillance Sciences Laboratory, Bedford Mass. Dealing with special systems, he has developed the Cartesian coordinate automatic tracking system (CARTRAC) for the Tactical Air Command, designed a semiautomatic intercept computer (SAINT), and, more recently, evolved techniques for detecting the launching of missiles and satellites.

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R. J. Slifka

Richard J. Slifka was born in Milwaukee, Wis., September 10, 1927. He received the B.S.E.E. degree from the University of Wisconsin, Madison, in 1951.

After graduation, he joined the AC Spark Plug Division of the General Motors Corp., Milwaukee, Wis., as a service engineer. In this capacity, he was associated with the Gunsight and Bombing Navigational Computer projects. In 1956, he joined the AC Spark Plug Engineering Department, in which his projects included test planning and coordination of field sites for the inertial guidance equipment for the Thor Missile. He is now a senior project engineer, responsible for field operations for the inertial guidance equipment for the Thor Missile.

Mr. Slifka is a member of the AIEE.

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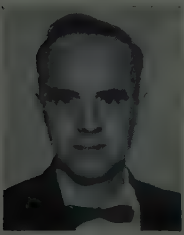
A. G. Smith

Alex G. Smith was born in Clarksburg, W. Va., on August 12, 1919. He received the B.S. degree in physics from the Massachusetts Institute of Technology, Cambridge, in 1943. During the remainder of the war, he worked at the M.I.T. Radiation Laboratory on magnetron development, and participated in the writing of a volume of the Radiation Laboratory Series. In 1948, he received the Ph.D degree in physics at Duke University, Durham, N. C., for which he wrote a dissertation on microwave spectroscopy.

Since that time he has been at the University of Florida in Gainesville, where he has held the ranks of assistant professor, associate professor, and, currently, full professor of physics. His research interests at the University have included microwave spectroscopy, military VHF electronics development, atmospheric optics, and low-frequency planetary radio astronomy.

Dr. Smith is a member of Sigma Xi, the American Physical Society, the American Optical Society, the American Astronomical Society, and the Astronomical Society of the Pacific.

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R. T. Smith

Robert T. Smith (M'56) was born in Kansas City, Mo., on October 25, 1930. He received the B.S.E.E. degree from the University of Kansas, Lawrence, in 1952, and while there he was employed by the University of Kansas Research Foundation on an analog computer development project.

Upon graduation, he joined the U. S. Naval Ordnance Test Station, China Lake, Calif., where he worked on air-to-ground fire control systems, special weapon fuzing systems, and automatic process control systems. In 1956, he assumed the responsibility of program manager for an airborne fire control program. He joined the Philco Western Development Laboratories, Palo Alto, Calif., in 1958, and is now manager of the advanced applications section of the satellite systems department. His current activities are in the analysis and design of current and proposed space electronic programs.

Mr. Smith is a member of Eta Kappa Nu and Sigma Tau.

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R. C. Spencer

Roy C. Spencer (M'46–SM'50–F'60) was born on April 14, 1901 in Pennellville, New York. He received the B.A. degree from Cornell University, Ithaca, N. Y. in 1922, and the Ph.D. degree in physics from Columbia University, New York, N. Y., in 1932.

From 1922–1923, he was a Fellow in Astronomy at Swarthmore, College, Swarthmore, Pa., and from 1923–1924, a Hechscher assistant in X rays at Cornell University. In 1926 and 1927, he was a Radio Tube Development engineer at the Westinghouse Lamp Co., Bloomfield, N. J. From 1927 to 1931, he was an instructor in physics at Columbia University, and from 1931 to 1941, he held positions of instructor, assistant professor and associate professor of physics at the University of Nebraska, Loncoln.

In 1941 he was granted a leave of absence from the University of Nebraska to become a staff member of the Radiation Laboratory of the Massachusetts Institute of Technology, Cambridge, Mass., where he carried out research on microwave antennas and diffraction theory. He was chief of the Antenna Laboratory of the Air Force Cambridge Research Center, Bedford, Mass., from 1946 to 1955, when he joined the Sylvania Missile Systems Laboratory, Waltham, Mass., as a senior engineering specialist and consultant. In 1958, he joined the Martin Company, Baltimore Division, Baltimore, Md., as principal staff scientist.

Dr. Spencer is a Fellow of the American Physical Society and the AAAS, and a member of the American Optical Society, the American Mathematical Society, Sigma Xi, and RESA (Research Society of America). He served on the U. S. National Committee of Commission VI of URSI and was a U. S. delegate to three International Assemblies in 1952, 1954, and 1957.

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J. O. Spriggs

James O. Spriggs (M'60) was born in Syracuse, N. Y., on May 18, 1914. He received the B.A. degree in sociology in 1936, the LL.B degree in 1940, and the B.A. degree in physics in 1945, all from Syracuse University, Syracuse, N. Y. He is a member of the New York State Bar.

From 1940 to 1942, he practiced law with Spriggs, Merry, and Hartman of Syracuse. He was engaged in research and development in the fields of direction finding and countermeasures at the Naval Research Laboratory, Washington, D. C., from 1942 to 1949, receiving the Meritorious Civilian Service Award in 1946. From 1949 to 1958, he was in the Office of the Secretary of Defense, first in the Research and Development Board, the Office of the Assistant Secretary for Research and Development, and then the Office of the Director of Guided Missiles. Since 1958, he has been on the staff of the Advanced Research Projects Agency in the ballistic missile defense and space fields as Special Assistant to the Chief Scientist, to the Deputy Director, and, most recently, as Program Manager for the Courier communications and the Transit navigation satellite project.

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R. E. Stalcup was born in Bloomfield, Ind., on February 19, 1927. He received the B.S.E.E. and M.S.E.E. degrees from Purdue University, Lafayette, Ind., in 1950 and 1959, respectively. He is now working on the

Ph.D. degree at the University of Illinois, Urbana.

R. E. Stalcup

After receiving the B.S.E.E. degree, he joined ITT Laboratories, Fort Wayne, Ind., where he has done research and development in the following fields: design and development of magnetic amplifiers for servo applications and for computing devices; design of radio and television circuitry; design and development of digital computer circuitry, a radar system test set and fuze test set; Terrier and Bomarc missile programs; analysis, design and development of aircraft fire-control systems; studies pertaining to the use of radiometers in guidance systems; and analysis of infrared attitude control system for space vehicle (orbiting) and studies of applicability of Doppler radar to terminal guidance problems of space vehicles.

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A. R. Tanguay

Armand R. Tanguay (S'49–A'51–SM'57) was born in Beloeil, Quebec, Can., on February 1, 1924. He received the B.S.E.E. degree from the University of Massachusetts, Amherst, in 1950 and the M.S.E.E. degree from the Massachusetts Institute of Technology, Cambridge, in 1951. While at M.I.T., he was a research assistant at the Digital Computer Laboratory (Project Whirlwind) doing research on electrostatic storage tubes.

From 1951 to 1957, he was employed at the Cornell Aeronautical Laboratory, Buffalo, N. Y. He was Associate Research Engineer from 1951 to 1955, concerned with research and development of military electronic systems and responsible for development of advanced missile ground guidance equipment. From 1955 to 1957, he was head of the Systems Analysis Section in the Weapons Systems Design Department, responsible for analysis, evaluation and data processing for the Lacrosse missile program, as well as for studies on advanced missile systems. Now with Radiation Incorporated, Orlando, Fla., he is Head of the Systems Research Department, with responsibility for advanced missile systems, space vehicle systems and guidance studies, computer simulation, and research in automation.

Mr. Tanguay is a member of Phi Kappa Phi and a senior member of the American Astronautical Society and Sigma Xi.

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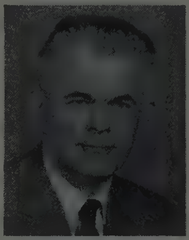
Frederick J. Tischer (SM'55) was born on March 14, 1913, in Plan, Austria. He received the Ph.D. degree from the University of Prague, Czechoslovakia, in 1938, and did advanced study at the University of Berlin, Germany.


F. J. Tischer

Before coming to the United States in 1954, he directed activities in microwaves at the Royal Institute of Technology, Stockholm, Sweden. He headed a research group working on electronic guidance at Redstone Arsenal, Huntsville, Ala., until 1956, when he joined the staff of the Electrical Engineering Department, The Ohio State University, Columbus. He is now an associate professor, and has taught, among other courses, one on space communications for two years. He is the author of "Microwave Measurements," and of numerous articles on communication theory, microwaves, and electromagnetics. He is the inventor of the Ring Resonator, the H-Guide, and several other microwave devices and antennas.

Dr. Tischer is a member of Sigma Xi.

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T. H. Vea

Theodore H. Vea (S'50–A'53–M'59) was born in Canton, Ohio on December 13, 1930. He received the B.S.E.E. degree from the University of Cincinnati, Cincinnati, Ohio, in 1953 where, on the cooperative engineering program, he was employed by the Wright Air Development Center in the search radar branch. In 1954–1955, he was an instructor in the electrical engineering department at the University of Pittsburgh, Pittsburgh, Pa., where he received the M.S. degree in 1955.

He has been employed by the Westinghouse Air Arm Division, Stanford Applied Electronics Laboratory, and the General Electric Computer Department where he has worked on airborne fire control systems, electronic countermeasure systems and the development of a data processing system for banking applications. In 1957, he returned to Stanford University to do graduate study and was a Research Assistant in the Radio Propagation Laboratory. In 1959, he joined the Philco Western Development Laboratory, Palo Alto, Calif., where he is employed as a project engineer in the advanced applications section of the satellite systems department. His activities include the analysis and design of communication systems for satellite applications.

Mr. Vea is a member of the American Institute of Physics, the Association for Computing Machinery, and AIEE.

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Joseph H. Vogelman (M'46–SM'49–F'59) was born in New York, N. Y., on August 18, 1920. He received the B.S. degree in 1940 from the College of the City of New York, N. Y., and the M.S. and Ph.D. degrees in 1948 and 1957, respectively, from the Polytechnic Institute of Brooklyn, N. Y., both in electrical engineering.

J. H. Vogelman

In 1945, after several years at the Signal Corps Radar Laboratory at Ft. Hancock and Belmar, N. J., he joined the staff of the Watson Laboratories, Red Bank, N. J., where he served as Chief of the Development Branch until 1951, responsible for research and development of test equipment and microwave components and techniques. In 1951, he moved to the Rome Air Development Center, Griffiss Air Force Base, Rome, N. Y. From 1951 to 1953, he was Chief Scientist in the General Engineering Laboratory and consultant on UHF and SHF theory and techniques to the Air Force. From 1953 to 1956, he was Chief of the Electronic Warfare Laboratory, directing all research and development in ground-based electronic warfare for the Air Force. From 1956 through June, 1959, he was Technical Director of the Communications Directorate with responsibility for the Air Force research and development effort in ground-based and ground-to-air communications. He is now Director of Research and Development at Dynamic Electronics Division, Capehart Corporation, Richmond Hill, N. Y.

Dr. Vogelman is a Fellow of the AAAS and a member of Eta Kappa Nu, Sigma Xi, the AIEE, and the Armed Forces Communications and Electronics Association.

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J. M. Walter, Jr.

James M. Walter, Jr. (S'54–M'56) was born on March 13, 1931 in Harrodsburg, Ky. He received the B.E.E. degree from the University of Florida, Gainesville, in 1956.

After serving with the U. S. Navy for four years as an aviation electronic technician, he joined Radiation, Inc., Melbourne, Fla., in 1956. His work there has included participation in low-level electronic subcommutation, 250-channel multiplexer, RADICON, RADIAN, AFMTC Timing Terminal Equipment, PDM to digital coder and PPM to digital coder developments. He has recently been engaged in the development of a 96-channel digital data acquisition system for NORAIR and the acquisition system reported on in this issue.

Mr. Walter is a member of Sigma Tau and Phi Eta Sigma.

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George C. Weiffenbach was born on June 20, 1921 in Newark, N. J. He attended Harvard University, Cambridge, Mass. where he obtained the B.A. degree in 1949, and also Catholic University, Washington, D. C., where he obtained the Ph.D. degree in experimental physics in 1958.

He joined the Applied Physics Laboratory of The Johns Hopkins University, Silver Spring, Md., in 1951 and is now a senior staff member in the field of experimental physics.

Dr. Weiffenbach is a member of Sigma Chi, the American Institute of Physics, the American Geophysical Union and the Philosophical Society of Washington, D. C.

G. C. Weiffenbach

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William E. Williams, Jr. (SM'55) was born in Hattiesburg, Miss , on April 6, 1921. He received the B.S. degree in physics in 1942 from Mississippi State College.

After graduation, he joined the staff of the National Bureau of Standards where he was engaged in circuit design work on proximity fuzes for bombs and rockets. After a tour of duty in the Army from 1944 to 1946, he returned to the Electronic Instrumentation Section, NBS, where his field of work was in the design of a variety of special-purpose instruments to measure pressure, liquid levels, vibration, and oil film thickness, as well as pulse generators and temperature control equipment. From 1953 to 1956, he was engaged in the design of aircraft flight simulators with primary responsibility for the simulated radar and electronic countermeasures systems. From 1956 to 1959, he was responsible for the design of missile telemetry systems for the Diamond Ordnance Fuze Laboratory. Since then, he has been staff assistant for telemetry in the Office of Space Flight Operations, National Aeronautics and Space Administration.

W. E. Williams, Jr.

Mr. Williams is an associate member of the Telemetry Working Group of the Inter-Range Instrumentation Group.

L. Bruce Wilner was born in Fargo, N. D., on August 7, 1931. He attended Yale University, New Haven, Conn., with the assistance of scholarships, and received the B.E. degree in 1953. He studied at Stanford University, Stanford, Calif., and received an M.S. degree in mechanical engineering in 1954.

He was employed by Los Alamos Scientific Laboratory for four years and participated in several weapons test operations. In 1958, he was employed by Lockheed Missiles and Space Division, Palo Alto, Calif., and is now working in the Instrumentation Research Group of the Communications and Controls Department at LMSD. His fields of activity have included nuclear component design, gas thermodynamics, and high-pressure technology, relaxation phenomena, and instrumentation for measurements in fluids.

L. B. Wilner

Mr. Wilner is an associate member of ASME.

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IRE Awards, 1960

Founders Award

Haraden Pratt

For outstanding contributions to The Institute of Radio Engineers through wise and courageous leadership in the planning and administration of technical developments which have greatly increased the impact of electronics on the public welfare.

Morris Liebmann Memorial Prize

J. A. Rajchman

For contributions to the development of magnetic devices for information processing.

Browder J. Thompson Memorial Prize

J. W. Gewartowski

For his paper entitled "Velocity and Current Distributions in the Spent Beam of the Backward-Wave Oscillator," which appeared in the October, 1958 issue of the IRE Transactions on Electron Devices.

Medal of Honor Award

HARRY NYQUIST

For fundamental contributions to a quantitative understanding of thermal noise, data transmission and negative feedback.

Harry Diamond Memorial Award

K. A. NORTON

For contributions to the understanding of radio wave propagation.

W. R. G. Baker Award

E. J. NALOS

For his paper entitled "A Hybrid Type Traveling-Wave Tube for High-Power Pulsed Amplification," which appeared in the July, 1958 issue of the IRE TRANSACTIONS ON ELECTRON DEVICES.

New Fellows

R. B. Adler

For contributions to engineering education and to research in electronics.

H. H. Aiken

For contributions to the development of computer science and technology.

H. O. G. Alfven

For contributions to the understanding of the properties of plasmas.

W. M. Bailey

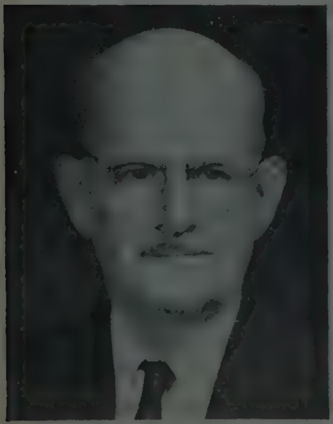
For contributions to capacitor technology.

Rudolf Bechmann

For contributions in the field of piezoelectric crystals.

J. I. Bohnert

For contributions in the field of microwave antennas.



J. F. Calvert

For contributions to electrical engineering and education.

D. K. Cheng

For contributions to engineering education and antenna theory.

Trevor Clark

For contributions to military electronics.

New Fellows

P. W. Crapuchettes

For contributions to microwave tube technology.

A. N. Curtiss

For contributions to radio and radar technology.

C. E. Dean

For contributions in the field of radio and television receivers.

G. A. Deschamps

For contributions to analysis of microwave components.

A. C. Dickieson

For contributions to wire and radio communications

Stephen Doba, Jr.

For contributions in the field of television signal transmission.

I. G. Easton

For contributions to impedance measuring techniques and equipment.

P. G. Edwards

For contributions in the field of telephone communication systems.

J. H. Felker

For contributions to computer technology.

New Fellows

G. J. FIEDLER

For contributions to electronic process control systems.

G. A. FOWLER

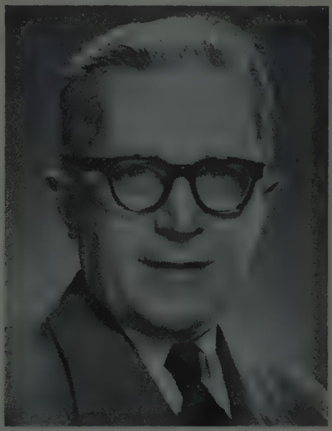
For contributions to atomic energy instrumentation.

D. W. FRY

For contributions to research in controlled fusion.

R. W. GILBERT

For application of electronics to measurement techniques.



G. S. GLINSKI

For contributions to the field of computation.

M. J. E. GOLAY

For contributions to the fields of communications and infrared technology.

GEORGES GOUDET

For contributions to engineering education and for the administration of research.

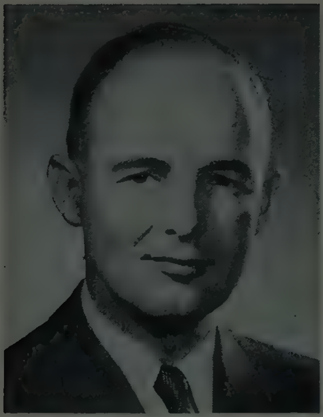
W. J. HAMM

For service in teaching and research.

H. C. HARDY

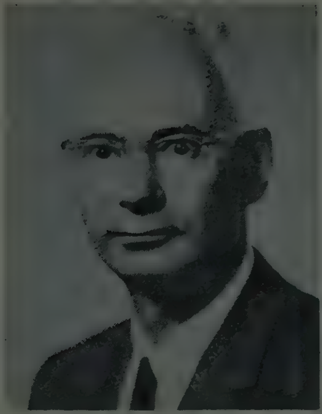
For contributions to electroacoustics.

New Fellows



R. A. Helliwell

For contributions to ionospheric radio propagation.



R. K. Hellmann

For work in military electronics.

S. W. Herwald

For contributions to servomechanism technology.

W. H. C. Higgins

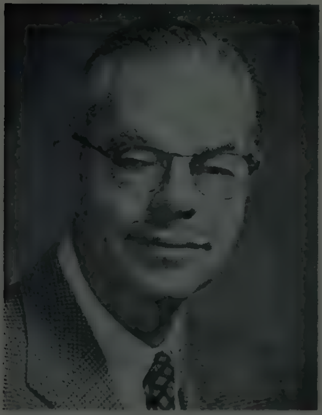
For contributions to military electronics.

C. L. Hogan

For pioneering in the application of ferrites.

J. M. Hollywood

For contributions to electronic countermeasures and color television.



F. L. Hopper

For contributions in underwater sound search and sound recordings

P. W. Howells

For contributions to color television.

H. R. Huntley

For contributions to communications engineering.

New Fellows

E. O. Johnson

For contributions to gaseous electronics and semiconductors.

Harwick Johnson

For contributions to the development of electron devices.

E. A. Keller

For contributions to sound recording and telephone switching systems.

C. L. Kober

For application of electronic techniques to defense systems.

Harry Krutter

For contributions to defense electronics.

J. G. Linvill

For contributions to network theory and transistor circuits.

W. A. Lynch

For contributions to engineering education.

J. M. Manley

For contributions to the theory of modulators and parametric amplifiers.

M. A. McLennan

For contributions in aero-medical electronics.

New Fellows

DOREN MITCHELL

For contributions to long-distance communications systems.

J. P. MOLNAR

For contributions to gaseous and solid-state electron devices.

L. H. MONTGOMERY, JR.

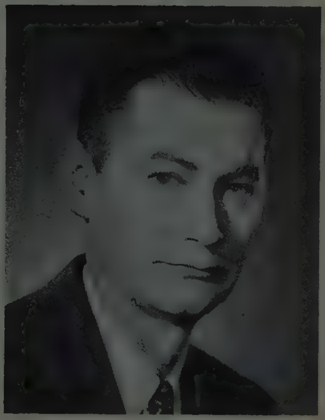
For contributions to medical electronics.

JOHN B. MOORE

For improvements in communications coding techniques.

THEODORE MORENO

For contributions in the field of microwave electronics.



T. H. MORRIN

For the administration of research in electronics.

SHOGO NAMBA

For contributions to the understanding of ionospheric radio propagation.

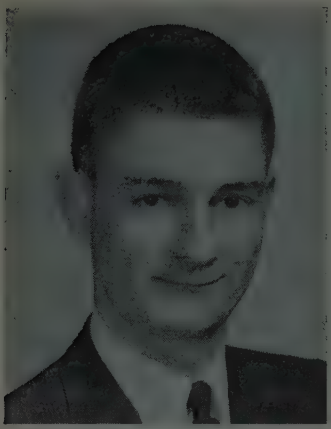
F. R. NORTON

For contributions to radar and television.

FRANZ OLLENDORFF

For contributions to electromagnetic theory and engineering education.

New Fellows



R. L. Pritchard

For contributions in the field of transistor circuits

R. L. Rau

For contributions to international radio communications systems.

L. L. Rauch

For contributions to the theory and practice of radio telemetry.

H. R. Reed

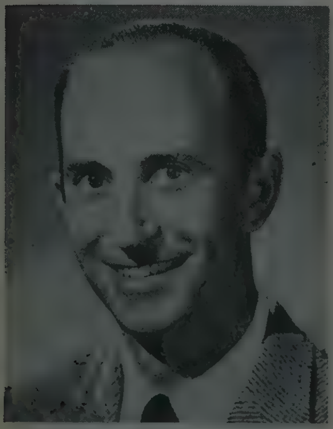
For contributions to engineering education.

V. C. Rideout

For contributions in the fields of microwave and computer education.

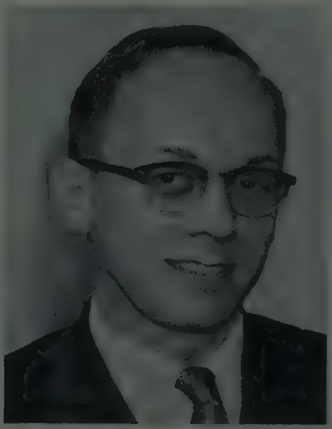
A W Rogers

For contributions in the development of electronic components.



V. H. Rumsey

For contributions to antenna theory and practice.



W T. Selsted

For contributions to the art of magnetic recording.

Samuel Sensiper

For contributions in the fields of microwave instrumentation and radiation.

New Fellows

William Sichak

For contributions to the techniques of microwave transmission.

R. L. Sink

For contributions to the field of digital instrumentation systems.

P. T. Smith

For contributions to the development of high-power transmitting tubes.

A. H. Sommer

For contributions in the field of photoemissive surfaces.

R. C. Spencer

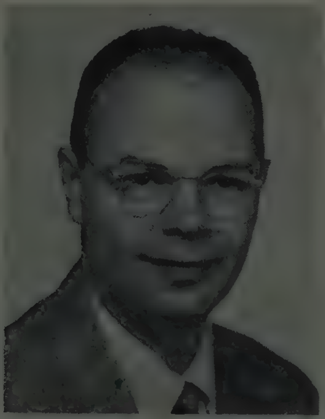
For contributions to the theory of microwave antennas.

W. E. Tolles

For the application of electronics to the field of medicine.

L. G. Trolese

For contributions to the understanding of radio wave propagation.



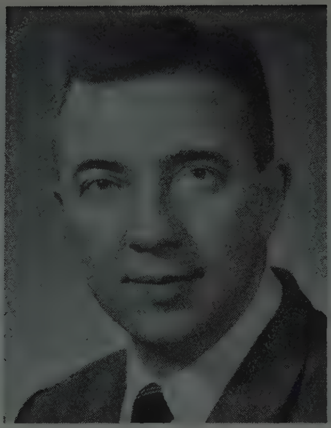
D. F. Tuttle, Jr.

For contributions to network theory and education.

G. E. Valley, Jr.

For contributions to military systems engineering.

New Fellows



J. A. VAN ALLEN

For the experimental discovery and exploration of radiation belts around the earth.

E. A. WALKER

For services as an engineering teacher and administrator.

W. M. WEBSTER, JR.

For contributions to gaseous electronic and solid-state devices.

LOUIS WEINBERG

For contributions to the field of network theory.

Books

Acoustique Musicale, Colloques Internationaux du Centre National de la Recherche Scientifique

Published (1959) by the Editions du Centre National de la Recherche Scientifique, 13, Quai Anatole France, Paris 7^{e}, France. 259 pages. 7×8¾. (In French.) 3.400 fr.

This volume contains the research papers presented at a three day international colloquium on musical sound, held at Marseille in 1958. One section of ten papers is about musical scales and harmony matters. The other section of thirteen papers is concerned with the physics of stringed and wind instruments. Electronics is involved only in the research instrumentation for spectrum analyses and frequency measurement. Nevertheless, an engineer interested in electronic means for music production, or for music reproduction, will find interesting and worthwhile basic information here about the traditional tone sources and the characteristics of their output waves. Among the musical instruments studied are the violin, cello, guitar, harmonium, pipe organ, French horn, trumpet, flute, clarinet, saxophone, bass horn, piano, and the human voice. The treatment of each instrument could not be comprehensive in one volume (or colloquium), but is illustrative and indicative of the possibilities for further research. The authors are from England, France, Germany, Hungary, Iran, the Netherlands, the USA, and the USSR.

Daniel W. Martin
The Baldwin Piano Company
Cincinnati, Ohio

Electron Tube Circuits, by Samuel Seely

Published (1958) by McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. 657 pages+14 index pages+21 appendix pages+xi pages. Illus. 6⅛×9¼. $10.50.

This book has been written for electrical engineering students in the fourth year of the electrical engineering curriculum. It consists of some 657 pages exclusive of appendixes and index and is to a large extent identical to the preceding volume (first edition). Some thirty-five pages of new material on semiconductors and transistor circuit applications have been added.

The author covers the same topics that were covered in the first edition with some change in the order and with the material on feedback organized in a separate chapter. There is also some reorganization in the chapters on oscillators and relaxation oscillators. New material has been added, in addition to the treatment of semiconductors and transistors, with respect to feedback and the effect of inverse feedback on the effective internal impedance of an amplifier. The discussion of internal impedance is considerably improved and expanded. The analysis of the cathode follower as a sample of the feedback circuit is very well done. A decided improvement has been made in Chapter 8 in the analysis of the difference amplifier. New material has also been included in the chapter on electronic computer circuits on the solution of simple differential equations. This chapter has 23 problems of which 10 are either identical with or slightly different from the problems in the old text.

Most of the new material is to be found in the expansion of the treatment of electron tube circuits. To a large extent the problems remain unchanged except for the addition of a few problems in most chapters, as for example Chapter 11, Tuned Potential Amplifiers, where the 29 problems include seven new problems added to the original 22 used in the corresponding chapter of the old text. A considerable improvement has been made in portions of the treatment of tuned power amplifiers.

The author makes a somewhat obvious effort to avoid the use of the word voltage, using potential instead. He also uses the term electron tube to refer to and to include all types of electron devices including transistors.

The author provides introductory treatment of the physical basis for semiconductor diode and transistor behavior as circuit components. The space provided for this treatment is not adequate for a satisfactory presentation. His development of the equivalent circuit of the transistor is related to the method used for tube equivalent circuits but is not so complete or satisfying. No mention is made of the h-parameters. Four terminal network theory is not adequately introduced and its usefulness is not made clear.

In summary, the second edition of this text provides a thorough and improved treatment of electron tube circuits. It does not, however, provide an integrated treatment of transistor and tube circuit theory with the proper distribution of emphasis in the semiconductor field.

E. M. Boone
Ohio State University
Columbus, Ohio

Information Transmission, Modulation, and Noise, by Mischa Schwartz

Published (1959) by McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. 454 pages+7 index pages+xv pages. Illus. 6¼×9¼. $11.00.

A concomitant of progress in any field is the gradual assimilation of previously esoteric material into earlier and earlier stages of the university curriculum in that field. Thus, in electrical engineering, the almost incomprehensible electromagnetic theory of eighty-five years ago is now meat for sophomores, and juniors and seniors are regularly exposed to the abstrusenesses of the modern form of network theory originating some thirty years ago. We are currently in the process of absorbing into the undergraduate electrical engineering curriculum the fruits of two decades of research into the application of statistical techniques to communication and control system design. The present book is a much-needed step in this process.

At present, a course in probability theory is, unfortunately, not generally required in the beginning years of an electrical engineering education—a situation which assuredly will change. Without the benefit of such a prerequisite course, however, and given only one semester in the senior year in which to teach communication theory, the question is: what should be taught? The answers to this have largely been the unsatisfactory extremes either of concentrating on classical probability theory and statistics, with perhaps some vague indication of the connection of all this abstract basic material to communication systems, or of concentrating on *ad hoc* systems, without giving an underlying statistical approach to the theory of communication systems in general. Professor Schwartz has now given us an excellent compromise solution.

In the first chapter, the author, by relying predominantly on the student's intuitive understanding of probability, gives a very brief introduction to the notions of information content and system capacity. He shows that two principal features of a communication system which limit its ability to transmit information by means of a physical signal are the maximum rate at which it will allow the signal to change, and the number of different signal states which the system can distinguish. The study of these features forms the unifying theme of the book. The first, related to the bandwidth of the system, is investigated in Chapter 2, in which are considered the Fourier analysis of signals and the effects of linear systems on signals. The second is related to the noise in the system, the nature and origins of which are studied in Chapter 5; here, for simplicity, noise is characterized only by its mean and mean-square values, which, defined as time averages, are easily related to the student's previous experience with time averages for nonrandom waveforms.

Chapters 3 and 4 form a detour into the detailed analysis of several conventional information transmission systems, both of the sine-wave (AM and FM) and pulse (PAM and PPM) modulation types. These chapters, interesting per se, are the more so for providing a background for Chapter 6, "Comparative Analysis of Information-Transmission Systems," which in many respects is the *pièce de résistance* of the book. Drawing together the results of the preceding five chapters, the author in this chapter compares the previously discussed conventional systems on the basis of the exchange relationship each exhibits between output signal-to-noise ratio and bandwidth. He shows that although some of these systems are more desirable on this basis than others, none approaches the performance predicted by information-theoretic considerations. This leads to a discussion of coded modulation schemes, and PCM in particular.

A lengthy final chapter is obviously—and both necessarily and regrettably—a postscript to the book, not intended for inclusion in the course. It is a superior, self-contained primer on the use in systems analysis of probability theory and statistics, the details of which had previously been avoided.

Professor Schwartz writes very lucidly indeed. He makes especially deft use of heuristic and intuitive arguments to focus attention on fundamentals and to clarify topics which are apt to be conceptually confusing for the beginner. Inevitably, there are a few minor features of organization with which this reviewer might disagree, and a modicum of those errors to which books, no matter how carefully written, seem to be prone. But in view of the over-all excellence of the book, to dwell on these would be mere carping. To those seeking a text which gives an elementary engineering insight into modern communication theory, this book is most highly recommended.

G. L. Turin
Hughes Research Labs.
Culver City, Calif.

Advances in Space Science, Volume I, Frederick I. Ordway, III, Ed.

Published (1959) by Academic Press, Inc., 111 Fifth Ave., N. Y. 3, N. Y. 382 pages+16 index pages +14 appendix pages+xii pages. Illus. 6¼×9¼. $12.00.

The book is a collection of six articles and an appendix by separate authors. These are:

1. Interplanetary Rocket Trajectories—by Derek F. Lawden
2. Interplanetary Communications—by J. R. Pierce and C. C. Cutler
3. Power Supplies for Orbital and Space Vehicles—by John H. Huth
4. Manned Space Cabin Systems—by Eugene B. Konecci
5. Radiation and Man in Space—by Herman J. Schaefer
6. Nutrition in Space Flight—by Robert G. Tischer

Appendix. A Decimal Classification System for Astronautics—by Heinz Hermann Koelle.

It is written for the space scientist who may be a specialist but desires to have a working knowledge of his associated fields. By existing standards it represents a rather thorough treatment of each of the six subjects presented. Much of the material in the first chapter was prepared as a report by Radiation Incorporated on an Air Force contract. Trajectories are considered mainly as a transfer between orbits which is generally predicated upon minimum propellant consumption and minimum time of flight criteria. The second chapter treats both the communications to other planets in space vehicles and to a satellite repeater, repeating signals back to the earth. The basic problems of the communications, the recent advances and the requirements of space systems are discussed in some detail. Typical systems are presented in terms of information bandwidth, transmitter power, antenna directivities, and receiver sensitivity. The third chapter treats power sources in the light of the space environment and the requirements of space vehicles. It considers both primary power sources and conversion techniques which include electrochemical systems, turbogenerators, magnetohydrodynamic conversion, thermocouples, thermionic conversion devices, quantum conversion techniques, (solar cells), and fission-electric conversions. Present state-of-the-art capabilities are outlined in some detail. In the fourth chapter many problems of the man in a space capsule are treated, including that of the closed life cycle. Specific methods of sustaining life in a space cabin are suggested and experiments are described. In "Radiation and Man in Space" an introduction is given on the effect of radiation upon human tissue. Then, the types and intensity of radiation near the earth and in space are treated in some detail. In the chapter "Nutrition in Space Flight" all problems of nutrition are treated from the standpoint of the closed cycle cabin system. Human needs for nutrition and human wastes are described, together with experiments which have been conducted in this field. In the Appendix, the classification adopted by the International Astronautical Federation is given under 90 General Headings and 900 Detailed Headings.

Conrad H. Hoeppner
Radiation, Inc.
Melbourne, Fla.

Our Sun, by Donald H. Menzel

Published (1959) by Harvard University Press, Cambridge 38, Mass. 340 pages+10 index pages. 201 figures. 6¼×9¼. $7.50.

This is the second and revised edition of the Harvard Astronomy Series book on the sun. The first edition appeared ten years ago. The text and illustrations are for the most part unchanged, although there are numerous revisions, modifications, and additions. In fact, there are twenty-three additional illustrations bringing the total to over 200. These include many recent photographs and spectra of the sun.

The author is now the Director of the Harvard Observatory and is widely known and respected for his outstanding theoretical contributions to astrophysics and for the impact that he has had on solar astronomy in this country.

The text is easy to read and can be understood throughout by an undergraduate student of engineering; it can be read with profit by any engineer interested in the scientific aspects of the sun. Nearly every aspect of solar observation is included: the early history of solar observations, the basic elements of chemistry, physics, spectroscopy, radio observations, ionospheric measurements, and atomic energy. The text concentrated principally on the description of the physical characteristics of the sun such as rotation, energy generation, sunspots, faculae, prominences, granules, flares, formation of solar absorption and emission lines, limb darkening, reversing layer, chromosphere, corona, spicules, solar magnetic fields, the relationship of solar activity to geomagnetic storms, auroras, solar cosmic rays, terrestrial weather, the uses of solar energy, and the relationship of the sun to the stars.

The instruments of optical observations and the techniques of solar measurements are simply described. The details of about 200 solar eclipses occurring in the twentieth century are given in groups according to each individual saros epoch. This is an interesting table to study in order to acquaint oneself with the patterns of solar eclipses.

This book is not a treatise, it does not use mathematics and there are no references to the literature. A large number of solar astronomers are mentioned, but the choice clearly reflects the author's personal approach. For example, Barbara Bell, a colleague, is referred to as often as are Professors Unsold and Minnaert combined.

Solar radio astronomy is sketchily presented and the material given doesn't properly represent the amount of effort in this field. No mention is made of the exciting work going on in Paris and Tokyo, for example.

This is an excellent introductory book on the sun for the engineer. It is lucidly written and well illustrated.

Fred T. Haddock
University of Michigan
Ann Arbor, Mich.

Digital and Sampled-Data Control Systems, by Julius T. Tou

Published (1959) by McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. 587 pages+7 index pages+11 bibliography pages+25 appendix pages. Illus. 6¼×9¼. $15.00.

In the field of sampled-data control systems a great amount of literature has appeared, but only a very few books have been published. This book takes its place along with the others as a notable contribution in bringing together the appropriate information. It is an excellent book.

The most notable features of this book are the logical development of the material, the manner in which various problems are faced up to, and the ease with which many of the topics are presented. The material reaches quite a bit beyond the theoretical and analytical aspects of sampled-data systems and touches on many of the practical problems such as quantization, analog-digital conversion, and specific circuit design considerations. In this respect one might ask why digital computer equipment and design was by-passed, but one readily visualizes the limitations on length of an already large volume.

In presenting the material the author assumes a strong reader background in control systems and servomechanisms. As a refresher, Chapter 2 summarizes this background material; but in an effort to condense, this chapter suffers from lack of clarity in places. The root-locus method for continuous data systems is probably the best part of this background material.

In dealing with sampling the author does not always distinguish between pulse and impulse modulations (*e.g.*, in the table of *z*-transforms, an impulse remains an impulse after impulse modulation), but here one can blame the negligence prevalent throughout the literature. What is nice in the exposition of both the methods and theories is the logical sequence in which problems are brought up and in the methods presented for their solutions. Somewhat too much emphasis is placed on Nyquist plots and their role in sampled-data analysis, but the use of root-locus and Bode plots is nicely stressed to balance the prevalent approaches in the literature. The theory of *z*-transformation is handled quite thoroughly and with considerable insight. System design, error analysis, and the use of error coefficients are treated. Statistical methods are used in several places and developed to at least a moderate extent.

The book shines not so much by its originality, but by the balanced bringing together of a wide variety of relevant methodology in a logical fashion. The amount of material presented is great, and the details are more often clarifying than cumbersome. It is clear that the author made a critical examination of the relevance and usefulness of the various methods, and in this reviewer's opinion he has exercised sound judgment in gaging their place and importance.

JOHN M. SALZER
Ramo-Wooldridge
Canoga Park, Calif.

Encyclopedic Dictionary of Electronics and Nuclear Engineering, R. I. Sarbacher, Ed.

Published (1959) by Prentice-Hall, Inc., 70 Fifth Ave., N. Y. 11, N. Y. 1417 pages+x pages. Illus. 7¼×10¼. $35.00.

This book is truly an encyclopedic work, embracing, as it does the electronic and nucleonic fields. It obviously represents a thorough scanning of the definitions promulgated by the leading military and professional sources and includes over 14,000 terms.

The dictionary has been written for engineers primarily, and particular effort has gone into the nuclear portion in an attempt to clarify nucleonic terms for the electronics engineer who wishes to enter this new field. Thus the reader will find many of the nuclear items expanded, with considerable explanatory matter.

There are profuse illustrations, including many drawings of various types of reactors, and typical schematic diagrams for various electronic circuits, also drawings illustrating construction of many common electronic components.

This reviewer has made a spot check of the material in several fields and is of the opinion that there is a wide variation in timeliness in these fields. In addition, this work suffers from the same fault that occurs so frequently where an author chooses to paraphrase existing standards. In many places the text gives credit to the IRE for a particular definition, yet there are various changes in the wording of the existing IRE Standard. While the basic meaning may not have been altered, the author has lost sight of the fact that these are no longer official IRE Standards and thus have no official authority.

The reviewer must also take exception to the statement on the dust cover "This is the ONLY single source to provide you with *all* the standard definitions approved by official technical societies." In view of the liberties taken with a large percentage of the IRE definitions checked this statement is obviously misleading.

To comment on specific sections: The nuclear portion was drawn previously from the ASME Glossary of Terms in Nuclear Science and Technology, interspersed with some IRE definitions (some altered) and others. Unfortunately, the above-mentioned Glossary is not widely accepted and was issued as a preliminary standard, pending a more definitive effort which is currently under way in the ASA. Thus this dictionary perpetuates a number of terms which are obviously inadequate. As an example, the various definitions of *Plateau* and related terms are restricted to Geiger counters, whereas in the IRE Standards these definitions apply to Radiation Counter Tubes, a much more inclusive category.

Some liberties have been taken with standard ASA and IRE symbology, *e.g.*, *T* is occasionally used for time, which should be designated by *t*, *T* being used for temperature.

In the semiconductor field, the book leaves much to be desired. There are numerous errors, *e.g.*, a diode is defined as "a two-electrode electron tube," although examples of semiconductor diodes are listed; in the definition of *diffusion length* the term *majority carriers* should be *minority carriers.* The transistor material is about five years out of date, all the equations being given in terms of the T-equivalent circuit, whereas today's texts and specification sheets use the hybrid *h* parameters almost exclusively.

Finally, the reviewer noted this amazing statement: "Standardization of transistor schematic symbols for point contact and junction transistors has not been attained." The text then proceeds to use both graphical and letter symbols at variance with the standardized IRE forms. The writer would have been better informed if he had consulted the IRE in this respect, as these items have been standardized since 1956 and 1957.

Despite these glaring errors, the book is still an important contribution to the art, provided the reader bears one thing firmly in mind: it is a dictionary, a glossary, definitely *not* a compendium of standard definitions, hence cannot be quoted or used with any official exactness. The reader who wants complete rigor in his definitions must refer to the official works of the respective technical societies. This particular society, the IRE, cannot give official sanction to the definitions ascribed to it, for the obvious reasons illustrated above.

R. F. SHEA
General Electric Co.
Knolls Atomic Power Lab.
Schenectady, N. Y.

Radar Meteorology, by Louis J. Batten

Published (1959) by The University of Chicago Press, Chicago 37, Ill. 154 pages+7 index pages+xi pages. Illus. 6¼×9¼. $6.00.

This book will serve to acquaint the radar designer with the pertinent characteristics of weather which effect the propagation of microwave energy through the atmosphere. Also, meteorologists who desire to know more about the use of radar as a meteorological tool will find this book useful. There is a wealth of references throughout the book, which enhances its usefulness to those doing research as well as to the radar designer and the synoptic meteorologist. The author states in the preface that his purpose is one of . . . "briefly describing in one small volume the progress made since the birth of radar meteorology" This he has done in a well organized manner.

The book starts with a short chapter on some of the basic principles of radar, apparently directed at the meteorologist. The following five chapters are devoted to a more detailed consideration of the properties and propagation of electromagnetic waves, the use of the radar equation for determining power returned from reflecting spherical particles, scattering from nonspherical particles and factors which influence the attenuation of electromagnetic waves.

Chapter seven covers the use of radar as a means of providing quantitative measurements of precipitation. Following this is a short discussion of the types of distortions which are to be expected as a result of having finite pulse lengths and beamwidths. The effects of antenna sidelobes are also mentioned.

With chapter nine the book begins to devote more to the meteorological aspects of "Radar Meteorology." It briefly describes mechanical, photographic and electronic methods of integration used in conjunction with rain gauges for determining the amount of rain falling over an area. The next chapter is concerned with the use of radar for cloud physics research. It covers such items as detection of precipitation in, and growth of, convective clouds, diameter and duration of convective echoes, the "bright band" phenomenon, precipitation generating levels and the use of radar for assessing results of cloud seeding.

The use of radar as a means of obtaining otherwise unavailable information on thunderstorms, squall lines and tornadoes is discussed in some detail in chapter eleven. This is followed by a discussion of the use of airborne radar in aircraft operations as an aid toward reducing flight hazards.

In recent years radar has proved to be valuable in the study and prediction of the behavior of hurricanes and other large weather systems. A discussion of the information obtained by radar from such weather phenomenon is covered in chapter thirteen. This is followed by a short chapter on radar echoes (angels) which may occur during the absence of precipitation. The book concludes with a description of special techniques, such as isoecho contouring and stepped grey scale, which may be used in conjunction with the existing radar circuits to enhance the usefulness of the radar set as a meteorological device.

While this is not a profound book, it will serve to acquaint the uninitiated with the possible uses of radar for meteorological purposes. The author has done an excellent job of providing references to books and papers that are pertinent to the subject of Radar Meteorology.

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Scanning the Transactions

Single-sideband goes underwater. SSB enthusiasts may be be interested to know that single-sideband suppressed-carrier transmission has proved to be the only effective system for underwater communication. FM is entirely unsuited to transmitting acoustic energy through water because phase disturbance of the wave components becomes so great that the essential characteristics of the signal are completely destroyed. Conventional AM is subject to undesirable fluctuations in the signal due principally to the presence of the carrier. Suppressed-carrier SSB has provided the most satisfactory answer. A mobile system recently designed for use between a hovering helicopter and a submarine had the following rather typical performance characteristics: a carrier frequency of 8 kc, modulation frequency of 200 to 3000 cps, transmitted power of 20 watts, and a range of 5,000 to 10,000 yards. An interesting feature of the receiver, one which simplified the design and minimized the problem of signal drift, was the use of a double heterodyning system for recovering the signal in place of the usual process of generating and reinserting the carrier into the incoming suppressed-carrier signal. (N. D. Miller, "An underwater communication system," IRE TRANS. ON COMMUNICATIONS SYSTEMS, December, 1959.)

A new look in phonograph needles may be in the offing. While the change may not be detectable to the eye, it will be to the ear. The tips of present day needles are in the shape of a hemisphere having a radius of 0.5 to 1.0 mil. Unfortunately this shape, while convenient to fabricate, does not correspond to the shape of the stylus used to cut records. This difference causes tracing distortion during playback, an effect which can be reduced by decreasing the radius of the needle tip, but only at the expense of increasing the wear and tear on the record. The new needle being proposed is in the shape of a pyramid, closely resembling the shape of the cutting stylus. This new shape, while more costly to produce, results in less distortion and background noise and longer life for recordings, both monaural and stereophonic. (C. D. O'Neal, "The pyramid stylus," IRE TRANS. ON AUDIO, November-December, 1959.)

Ultrasonic welding celebrates its tenth birthday this year. It is still a poorly understood process, although ultrasonic welders have been successfully designed by trial and error methods for limited applications. Until the basic principles of the process are better known, however, the development of equipment will continue to be empirical and the limitations of the process will remain unknown. What is believed to be the first fundamental study of the mechanisms involved in ultrasonic welding has now been reported. It appears that the welding process, in which surfaces subjected to a clamping action are welded when made to slide in contact with each other, is directly related to the well-known frictional phenomena of galling and seizing. This study also disclosed how clamping force, duration of weld cycle, ultrasonic motion, and temperature are related to weld strength. Present indications are that practical ultrasonic butt, seam, and spot welders can be designed for welding thin-gauge sheets or small components of ductile materials. (J. N. Antonevich, "Ultrasonic welding equipment," IRE TRANS. ON ULTRASONICS ENGINEERING, February, 1960).

Choosing the best type of communication system for a particular application is neither a purely technical problem nor a purely economic one, but a complex combination of these two factors. One engineer with considerable experience in this field has devised a novel formula for rating one system against another which takes both factors into account. This rating formula provides a good first-order approximation of the relative merits of various systems, making it possible to narrow down the number of systems that must be analyzed in greater detail. The formula for system performance rating is given by

$$\text{SPR} = \frac{\text{number of repeaters} \times \text{error rate}}{10^8 \times \text{available bandwidth (mc)} \times \text{availability}} \times \log_{10} (\text{TPR} \times \text{RPR} \times 10^7),$$

where TPR (transmitter power rating) is equal to transmitter power × antenna gain/bandwidth (cps); RPR (receiver performance rating) is equal to receiver antenna gain/[receiver noise figure] × [bandwidth (cps)]; and "availability" is the fraction of the time that the transmission medium is available for use. The lower the system performance rating is, the more desirable it is, *i.e.*, the lower the cost is. It will be interesting to see how well this evaluation technique, which thus far has checked out well, stands the test of time and experience. (J. H. Vogelman, "Comparative evaluation of communications transmission media," IRE TRANS. ON COMMUNICATIONS SYSTEMS, December, 1959.)

The radiation tolerance of electronic components is receiving a growing amount of attention in the design of equipment for both civilian and military applications. A number of organizations have been exploring the effects of radiation on standard components for the past several years. The results of these investigations, reported in over 600 scattered documents, have now been compiled in a single information file. By means of this file it has been found possible to modify existing equipment by selective replacement of radiation-sensitive parts with standard parts having a higher radiation tolerance. With this approach, circuits remain essentially unchanged, yet the amount of radiation the equipment can withstand has been raised orders of magnitude. Thus while more radiation data is still needed, it appears we have now reached the point where radiant-tolerant equipment can be successfully built with what is presently known, without need for extensive radiation testing. (J. R. Burnett, "Radiation-tolerant electronic equipment," IRE TRANS. ON NUCLEAR SCIENCE, December, 1959.)

The development of new low-noise amplifiers has changed the emphasis in communication system design from considerations of receiver noise to considerations of external noise. The noise level of a maser can be as little as one-thousandth that of a conventional amplifier. The question arises how much of this improvement is actually realized in an operating system and how much is lost to external noise. The answer has now been worked out at least for one type of system—a tropospheric scatter system operating in the 1000 to 2000 mc range. It has been found that the cooled-crystal technique will provide a system improvement of 3 db, the parametric amplifier 9.5 db, and the maser 12 db. When the problems associated with liquid helium operation of masers are taken into account, it appears that the parametric amplifier is the best choice. (A. Feiner and D. Savage, "The effects of low-noise techniques on tropospheric scatter communications," IRE TRANS. ON COMMUNICATIONS SYSTEMS, December, 1959.)

Abstracts of IRE Transactions

The following issues of TRANSACTIONS have recently been published, and are now available from the Institute of Radio Engineers, Inc., 1 East 79th Street, New York 21, N. Y. at the following prices. The contents of each issue and, where available, abstracts of technical papers are given below.

Sponsoring Group	Publication	Group Members	IRE Members	Non-Members*
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Aeronautical and Navigational Electronics

Vol. ANE-6, No. 4, December, 1959

Frontispiece (p. 210)

Airborne Dual Antenna System for Aerial Navigation—W. Spanos and J. M. Ashbrook (p. 211)

This paper describes a 1000-mc dual antenna system which uses parallel-driven sector antennas. Methods for determining the performance in pattern interference regions are given together with applications to DME, Radar Safety Beacon and TACAN navigation systems. A flyable-model dual antenna system for Constellation and DC-6 type aircraft is described. Provision of an RF hybrid permits the simultaneous operation of two navigation equipments, such as DME and Radar Safety Beacon, from the same antenna system. A prototype of this antenna system on a DC-3 has provided improved performance for the TACAN navigation system. Flight tests with an experimental dual antenna system have shown improved performance for DME and Radar Safety Beacon systems. The results of flight tests show agreement with theoretically determined values of performance.

The Indeterminacy of Measurements Performed by Radar Equipment—R. Madden (p. 219)

A theoretical derivation of the indeterminacies of simultaneous position and velocity measurements of a reflector when using reflected electromagnetic waves is given. It is shown that the product of the indeterminacies is given by $1/(8\pi)/\lambda_g c$ where λ_g is the illuminating wavelength and c the speed of light. It is shown that the necessary consequence of nonsimultaneous measurements is an uncertainty as to whether the measurements are common to the same reflector. If this uncertainty is to be overcome, the reflectors must be spaced at a distance $\frac{1}{2}c\Delta T$ where ΔT is the time separation of the position and velocity measurement.

Principles of Electronic Navigation Systems—P. C. Sandretto (p. 221)

Electronic navigation systems of all types are discussed, and classified as classical or self-contained. The error rates and accuracies of the various systems are discussed.

Anticipatory Display Design Through the Use of an Analog Computer—L. J. Fogel and M. Dwonczyk (p. 228)

Modern high-performance aircraft currently are pressing the limitations of the human operator. The increased speeds compress the allowable reaction time to such levels wherein logical decisions, and even conditioned reflex actions, may no longer be possible. The only way to overcome this human limitation of manned aircraft performance is through the incorporation of anticipatory displays; displays which offer a prediction of the various parameters so that the human operator is projected ahead of the system. An aircraft was analog computer simulated, data reduction was programmed and the same computer was used to allow biophysical measurement, which furnished correlative measure. The effectiveness of various piloting techniques, as well as prediction intervals, was explored. The results indicated a first approximation to the design of improved displays through the use of anticipatory information.

PGANE News (p. 240)

Contributors (p. 242)

Annual Index, 1959 (follows p. 242)

Audio

Vol. AU-7, No. 6, November/December, 1959

The Editor's Corner (p. 137)

PGA News (p. 139)

The Pyramid Stylus—C. D. O'Neal (p. 140)

A new shape for a phonograph needle stylus is described which greatly improves the reproduction possibilities from phonograph records. The new stylus was specifically produced for use with 45°–45° stereo recordings, but will perform equally well on all microgrooved monaural recordings. The scheme described evolves a shape of reproducing stylus having a basic geometry relating to the cutting stylus used in forming all microgrooves. Computations, experimental data and charts are also used to support the improvements to be expected in reproducing reproduction performance.

Time-Frequency Scanning in Narrow-Band Speech Transmission—D. L. Subrahmanyam and G. E. Peterson (p. 148)

Two basic sampling methods for the transmission of intelligible speech with reduced channel capacity have previously been studied. The channel vocoder developed by Dudley is based on frequency band separation or quantization, and Schisser, and Fairbanks, Everitt, and Jaeger have developed a technique of time sampling. Sampling in both time and frequency offers a third major possibility which should be investigated.

For this study a narrow-band speech transmission system was constructed which scans the time-frequency plane. The system scans in either a sinusoidal or in a sawtooth manner over a frequency range of 200 to 7000 cps. In order to achieve the scanning, a variable carrier frequency oscillator, having a frequency shift of ±30 per cent of the carrier frequency and an amplitude modulation of ±2 db, has been developed. The process of scanning leaves empty spaces in the time-frequency plane, and a four-channel time-delay system which employs the method of dielectric recording has been used to fill the gaps in the time-frequency plane by repeating the signal samples.

When signal reiteration is employed with this system, a score of 75 per cent of monosyllabic phonetically balanced word lists was obtained with a scanning filter of 1000-cps bandwidth and a sinusoidal scanning rate of 30 times per second. This intelligibility is appreciably higher than that which can be achieved with a fixed 1000-cycle filter located in the frequency region of maximum intelligibility.

Circuits for Three-Channel Stereophonic Playback Derived from Two Sound Tracks—Paul W. Klipsch (p. 161)

Derivation of a third playback channel from two stereo sound tracks may be accomplished by several means. The center channel may be derived by recombination prior to power amplification; acoustically, after amplification, by using two center speakers; and in a variety of phase relationships, including the limiting case of equal signals (monophonic) in which either sum or difference combination may be chosen by polarity selection.

Characteristics of Degenerative Amplifiers Having a Base-Emitter Shunt Impedance—William D. Roehr (p. 165)

In amplifiers having emitter degeneration, an impedance is sometimes used between base and emitter. A common case occurs when several emitter followers are used in cascade. The resistors become necessary in order to provide some measure of stability.

In an audio amplifier of similar design it became necessary to know what effect this shunt resistor would have upon the input impedance of the stage. As a result, the analysis given in this paper was performed. Experimental measurements which support the resulting equation are given.

This analysis and these measurements led to the discovery of a circuit which would exhibit a high ac input impedance, yet the resistors in the dc base circuit could be kept low to provide good stability. Although it can be shown that circuit power gain is the same regardless of whether a resistor is used in series with the emitter or the base to obtain a high input impedance, considerations of dc stability and distortion demand a more thorough investigation of this problem.

Design principles are outlined for this high input impedance stage and an example is worked out in detail. Supporting measurements are given.

Correspondence (p. 169)

Contributors (p. 170)

Annual Index, 1959 (p. 172)

Communications Systems

Vol. CS-7, No. 4, December, 1959

Frontispiece and Guest Editorial (p. 227)

Comparative Evaluation of Communications Transmission Media—Joseph H. Vogelman (p. 230)

The selection of communications transmission media for any specified operation is neither a purely technical problem nor a purely economic one, but a complex combination of these two factors. To minimize work involved in evaluating all the possible choices, a first-order approximation method is provided to limit the number of specific transmission media which must be considered in any specific application. This method provides a numerical measure of the expected combination of technical and economic factors and is based on the author's empirically-derived experience with communications equipments and systems.

Design Considerations for Space Communication—J. E. Bartow, *et al.* (p. 232)

With the advent of Russian and American artificial earth satellites, the use of such vehicles for communication purposes has been the subject of considerable study both by military and commercial organizations in the communications field. The purpose of this paper is to delineate the problems involved in space communication, the assumptions that must be made and the technical limitations which determine the communication system that should be used for a particular time frame. Some characteristics for an optimum system are stated. Some technical characteristics of the first successful satellite communication system are given.

Development Trends in USAF Global Communication Systems—C. A. Strom, Jr. and A. A. Kunze (p. 241)

This paper briefly summarizes one initial development approach for the 1962–1965 USAF Global Communications Systems. A brief review is given of some of the early communications systems work and the present design philosophy for the 600-voice bandwidth channel, 6000-mile trunk system is discussed. Specific requirements of the ground-based Air Force common-user system are listed and developments intended to provide much of the equipment for this time period are noted. This paper concludes that by adopting the system approach, the Air Force will be provided with a greatly increased communications capability at a total cost which differs little from today's annual expenditure.

An Underwater Communication System—Norman D. Miller (p. 249)

Communications between submerged submarines and surface craft require energy to be propagated through sea water. The only practical method for transmitting this energy is by acoustic means. Both FM and AM transmission systems have been used, but for long-range communication, a single-sideband suppressed-carrier transmission system is the only effective way of propagating sound energy. The requirements for the transmission system and diagrams of the circuitry are given.

Wide-Band Facsimile Transmission Over a 900-Mile Path Utilizing Meteor Ionization—W. H. Bliss, *et al.* (p. 252)

Preliminary tests of facsimile transmission over a 910-mile path have been made at 40 mc by means of intermittent meteor ionization. Printed matter was scanned at a rate of 2 frames per second, with a resolution of 67 elements per inch. An all-electronic facsimile system recorded a picture when the received signal rose above a preset threshold.

Tests were conducted employing bandwidths up to 110 kc with keying frequencies up to 73 kc, still maintaining the same optical resolution in the facsimile material. With the equipment used in the tests, it appeared that the maximum bit rate for acceptable facsimile was about 48,000 bits per second at 2 per cent duty cycle and about 86,000 bits per second at 1 per cent duty cycle. Examples of signal recordings with received facsimile are shown.

Preparations are being made to conduct similar tests, employing a higher information rate and wider bandwidth, on a frequency near 50 mc.

A Very High-Speed Facsimile Recorder—G. M. Stamps and H. C. Ressler (p. 257)

An operating facsimile system is described which is capable of transmitting black and white graphic information at the rate of 24 inches of copy feed per second. The system uses conventional line-by-line sequential scanning at the transmitter and a multistylus recorder with 100 lines per inch resolution at the receiver. The electrolytic recording medium employed is characterized by a relatively low internal impedance, which permits close spacing of the stylus electrodes and the use of direct-coupled transistors for the individual marking amplifiers. The high transmission speed results from the use of individual stylus electrodes for each elemental area along the scanning line and the holding of the video marking signal at each stylus for time intervals approximating the duration of a complete scanning line. The system described uses a 100-stylus block assembly, restricting the length of the scanning line to one inch. One thousand-stylus block assemblies have been made and these will permit reproduction of copy 10 inches wide.

The NBS Meteor Burst Communication System—R. J. Carpenter and G. R. Ochs (p. 263)

This project was undertaken in 1955 to investigate the properties of the intermittent reception of VHF signals over long distances by meteoric propagation and their communication usefulness. To accomplish this, a complete duplex teletype system operating at about 50 mc was constructed, and results are reported from tests made over a 1277-km east-west path. Comparisons are made of burst transmissions of 10, 20, 40, and 80 times normal teletype speed and of variations in a number of control system parameters.

For the system under test, the optimum speedup was 40×, which produced a daily average channel capacity of about 40 wpm with a character error rate of about 0.0035 (with the best control system settings). Higher speedup ratios are advocated for future systems. The most serious causes of outages in this type of system are atmospheric noise and sustained multipath distortion from competing modes such as E_s and auroral propagation.

White Alice System—Design and Performance—A. L. Durkee, *et al.* (p. 272)

White Alice is an extensive communications network covering the entire state of Alaska. It provides essential defense communications as well as service for other government agencies and civilian commercial service. Both telephone and telegraph channels are provided with capacity for future growth. The Western Electric Company, as prime contractor for the USAF AMC, undertook the engineering, procurement, and installation of the electronic equipment as well as construction of some of the sites.

The Bell Telephone Laboratories participated in the basic systems engineering and furnished consulting assistance in new scientific areas. New designs were made or supervised by the Laboratories. The White Alice system plan is based on the study by the Bell System of the communications needs of the area.

Radio transmission was chosen for the majority of the routes. Wire and cable were considered either unreliable or impractical in the Alaskan terrain and climate. Where several hundred voice channels were required, microwave line-of-sight radio was used. For the majority of the links, tropospheric beyond-horizon transmission was used because of the economy and simplification of maintenance and logistic problems.

The transmission objectives for the entire system were a median noise in the worst month of 38 dba and a maximum of 1 per cent outage in the worst month. These objectives were for 4000-mile circuits and were prorated to each link.

The microwave portions of the system were engineered in the same way that has been used previously. Tropospheric beyond-horizon was a relatively new mode of radio transmission and the engineering procedures were less well defined. Propagation experience from a year-long test in Newfoundland, in addition to research by other groups, was used as the basis for the system engineering. These data include variation of signal strength from day to day and seasonally, the effective gain of the large antennas employed, the characteristics of the rapid fading typical of tropospheric propagation, and the effects of geographical profile and separation of the stations.

Towards the completion of the project, comprehensive transmission tests were made of many of the links of the system to determine that the design objectives had been met. In particular, a 2400-mile circuit traversing much of the system was tested. The transmission stability and noise performance over a 30-day period was well within the limits set in the objectives. Teletype and voice performance were eminently satisfactory.

White Alice has been operating continuously since it was turned over to the Air Force in March, 1958. Its reliability has been gratifying to the military and other agencies using it. It is certain to be a significant factor in the development of this last frontier.

Installation and Operational Aspects of a Private Television Microwave System—Aaron Shelton (p. 278)

The problems of installation, operation and maintenance of a 165-mile, six-hop, private, intercity television relay system are reviewed. Careful planning of the path to be used is required. Equipment, its housing, power and frequency chosen for this service require careful evaluation. Emergency power supplies are required. Numerous modifications to commercial equipments were found necessary. The problems of sound diplexing through a multihop system proved difficult, but possible and worthwhile. Careful maintenance schedules and techniques developed for checking operation of each station proved their value. Communications for a multihop system become very important. Manpower required for a system of this magnitude proved to be greater than expected. Equipments designed and field checked for multihop installations can only be considered if truly dependable service is required and expected.

Practical Considerations in the Design of Minimum Bandwidth Digital Frequency Modulation Systems Using Gaussian Filtering—W. L. Glomb (p. 284)

The characteristics of a digital frequency modulation system are reviewed. The Gaussian frequency response is specified and optimum filters for transmitter and receiver (both predetection and postdetection) are developed. The relationship between peak deviation and system threshold is explored and an optimum deviation established. The location of the Gaussian filters in the system is discussed. It is shown that for best threshold performance, all receiver filtering should be predetection filtering and that optimum peak deviations under these conditions may correspond to the −4-db point of the IF filter. Practical details of transmitter and receiver design are discussed.

These results have been applied to a 2000-mc communication equipment.

A Study of the Technical and Economic Feasibility for Tropospheric Scatter Circuits in Primary Toll Networks of Underdeveloped Countries—C. A. Parry (p. 290)

The inherent capabilities of the tropospheric scatter circuit have emphasized the need for careful economic and technical evaluation. It is suggested that this should be carried out in two phases. The first is a broad study which determines the essential system parameters for which detailed analysis is necessary. The second phase is concerned with this detail. Outlines are given relevant to the principal factors associated with the first of these phases. These cover such areas as trunk circuit growth, required channel, capacity at the end of the amortization period, revenue potential, optimum routing, choice of transmission method, and the optimum number of tandem tropospheric scatter links. The performance sacrifice for a minimal cost "first in" system and the manner in which design accuracy may be traded against probable propagation reliability are two of the problems discussed. Evaluation of the scatter link may be based on a performance index and data relevant to this are given. It is concluded that adequate analysis of the modern system capability requires a sophisticated engineering approach for which the present paper is a guide.

50-kw Antenna Switching System—John W. Smith (p. 295)

There are many HF radio installations today that illustrate the need for an improved antenna switching technique. Today there is a growing requirement to connect rapidly any one of several 50-kw transmitters to any one of many antennas. An antenna switching system must be broad-band, convenient to use, fast, safe for operating personnel, and free of appreciable impedance discontinuities. This paper describes several components and techniques that solve the transmitter-antenna selection problem up to a 50-kw power level. The influence of switching techniques on RF transmission, impedance conversion and antennas are also discussed.

The Effects of Low-Noise Techniques on Tropospheric Scatter Communications—A. Feiner and D. Savage (p. 302)

This paper defines the expected benefits to be derived from the use of low-noise receiver techniques when applied to tropospheric scatter systems in the 1000- to 2000-mc range. Calculations were made of the expected antenna noise temperatures for "smooth" and "rough" ground locations. These data, plus the known and anticipated performance of various low-noise techniques, permitted the calculation of increased sensitivity. The sensitivity improvement is then expressed in terms of increased range, decreased transmitter power, and decreased antenna size.

Contributors (p. 307)

Education

Vol. E-3, No. 1, February, 1960

Editorial (p. 1)

The Research Flavor—John S. McNown (p. 2)

Research should be made a significant part of undergraduate engineering education in order to orient the student toward the solving of new problems. He should learn of the frontiers of his field of study, and should know of the people who are pushing back these frontiers.

An increase is needed in the number of courses designed for guiding a student to define problems, to make assumptions, to search for his own methods of solving them, and to develop a way to report his findings. Almost as a dividend, new ideas and the exercise of originality provide a strong motivation to learning.

The Scientific Manpower Shortage and Its Implications for the Engineering Technician—John Merrill (p. 3).

Some of the background of the present scientific manpower shortage is reviewed. Utilization as an important aspect of engineer efficiency is discussed. With this setting, the engineering technician's complementary function with the engineer is introduced. The limited availability of technicians is cited. A discussion of the *stepping-stone* aspect of the technician job, an area worth consideration because of its ramifications, is presented and the concept of a four-year program for engineering technicians is shown to have significant value in coping with several phases of technician training.

Four areas of study and investigation are suggested that can contribute to solution of the over-all need of increased scientific manpower through increased utilization. Utilization would be improved by the proper preparation of an adequate number of engineering technicians.

Circuit Analysis—A Problem in Semantics—Thomas M. Adams (p. 10)

The teaching-learning process in circuits is enormously complicated by the multiple meanings and shades of meanings which are inherent in most of the technical terms. A method is offered for reducing every term relating to "functions" to its basic meaning for the particular circuit application under study. Using this method, which might be called "current analysis," it becomes possible to explain detailed circuit phenomena to students with little or no technical or mathematical background.

A General Course in Traveling Waves—R. K. Moore (p. 15)

Traveling waves must be considered whenever the time for a disturbance to propagate from one place to another is important. A course with this as its theme has been developed to replace the usual course in transmission lines or distributed constant circuits. Emphasis is placed on the transmission line as a teaching vehicle, but plane waves in space; vibrating strings and membranes; acoustic waves in gases, liquids and solids; heat conduction; and chemica diffusion are also treated. The analogies between the "telegraphers' equations" and comparable equations describing nonelectrical phenomena are stressed.

Transients are treated for two special cases: 1) lossless lines and acoustics, and 2) diffusion and heat transfer. Steady state analysis using phasors and the Smith Chart is applied to plane electromagnetic and acoustic waves at all angles of incidence, as well as the transmission line. Spherical acoustic waves are treated briefly.

A First Course in Electrical Engineering—P. R. Clement and W. C. Johnson (p. 19)

This article describes the introductory two-semester course in electrical engineering that has been developed over a period of five years at Princeton University. The aims of the course are to provide an introduction to the field of electrical engineering via the basic concepts of electricity and magnetism (which are developed through Maxwell's equations in the integral form), to apply these concepts in developing the fundamentals of energy conversion and circuit theory, and to carry forward, in a continuous and integrated way, a modern treatment of network analysis. Thus, the treatment proceeds from field ideas to circuits and physical apparatus, and to their mathematical models. With the circuit relations formulated, attention turns to the analysis of networks, starting with network topology and extending through pole-zero ideas. The treatment stops just short of the Laplace transform. This course is intended to serve as a foundation for subsequent courses such as electronic circuits, energy conversion, and advanced network analysis and synthesis, and has been developed with the purpose of providing the student with a unifying point of view for these varied topics.

The Organization of a Laboratory Electronics Course—D. Blackwood (p. 22)

This paper describes the organization of a Laboratory Electronics Course in the Department of Aircraft Electrical Engineering at the College of Aeronautics, Cranfield, England. The course caters to two standards of specialization and to students with a wide range of previous training.

The minor specialization course concentrates solely on fundamentals and the basic concepts, while the major course is carried to the threshold of design procedures. The contents of both courses are limited to the student's previous training and the time available for practical work. The over-all standard of the major course is that of the M.Sc. degree of a British University.

Four different ways of construction and presentation of experiments are described and the limitations of each are discussed. Finally, the laboratory layout, equipment, and scheme of work for both courses is described.

A Supplementary DC Machine Test for Sophomores—E. Della Torre and C. V. Longo (p. 25)

A blocked-rotor test may be used to determine curves for a family of dynamic operating characteristics for dc machines. Conventional performance characteristics may be predicted from these curves. The authors have used this test procedure to supplement the usual tests on machinery in the sophomore laboratory. The new procedure emphasizes the fundamental principles of machine operation and also points up the importance of certain details which have a significant effect upon performance. The test would logically precede the usual load tests. The latter are then anticipated with enthusiasm since they provide confirmation of performance predicted from the static tests.

Contributors (p. 29)

Information Theory

Vol. IT-5, No. 4, December, 1959

Frontispiece (p. 148)

Editorial (p. 149)

A Class of Systematic Codes for Non-Independent Errors—N. M. Abramson (p. 150)

A class of systematic codes has been obtained which will correct all single errors and all double errors which occur in adjacent digits. These codes use significantly fewer checking digits than codes which correct *all* double errors. In addition, because of inherent regularities in their structure, these codes may be instrumented in a strikingly simple fashion.

A Probabilistic Model for Run-Length Coding of Pictures—Jack Capon (p. 157)

A first-order Markoff process representation for pictures is proposed in order to study the picture coding system known as run-length coding (differential-coordinate encoding). A lower bound for the saving in channel capacity is calculated on the basis of this model, and is compared with the results obtained by previous investigators. In addition, this representation is shown to yield an insight into the run-length coding system which might not otherwise be obtained. The application of this probabilistic model to an "elastic" system of run-length coding is also discussed.

A Note on Invariant Relations for Ambiguity and Distance Functions—C. A. Stutt (p. 164)

Woodward's result for the ambiguity function, that the volume associated with its squared magnitude over the time-shift and frequency-shift plane is a constant, has been shown to be true also for a cross-ambiguity function for two time functions. If complex time functions have been obtained by means of a Hilbert transformation from real time functions, it is found for the cross-ambiguity function that the volumes under the squared real part and under the squared imaginary part are constant and contribute equally to the volume under the squared magnitude function. A "distance" function for two time functions is defined to be the integrated squared difference between these functions. The relation for the squared real part of the ambiguity function readily yields an invariant relation for the volume associated with this distance function in the case of Hilbert-derived complex time functions. An especially simple invariant relation for the "mean" distance, as computed over the time-shift and frequency-shift plane, exists for such time functions having finite energy and finite mean value.

On Upper Bounds for Error Detecting and Error Correcting Codes of Finite Length—Nelson Wax (p. 168)

Upper bounds for error detecting and error correcting codes are obtained in this paper. One upper bound is found by exploiting the geometrical model of coding introduced by

Hamming. The volume of an appropriate geometrical body is compared with the volume of the unit cube, in getting the first upper bound. An improvement on this upper bound can be found by introducing a mass density function, and comparing the mass of the body with the mass of the unit cube.

A comparison is made with known upper bounds, and with best codes found thus far. The improved upper bound given here is frequently somewhat smaller than previously known upper bounds.

The Probability Density of the Output of an RC Filter When the Input Is a Binary Random Process—H. A. McFadden (p. 174)

A new method is given for obtaining the probability density of the output of an RC filter when the input is a stationary binary random process. The axis-crossing intervals of the input are assumed to be statistically independent and identically distributed, but with an arbitrary density function. The method involves a linear integral equation which can be reduced by Laplace transforms. A new family of solutions is given which includes two previously known cases: the random square wave of Poisson type, and the periodic square wave with random time origin. The general result of this family is given in terms of tabulated functions. For other solutions, a recursive technique may be necessary.

Recurrent Events in a Bernoulli Sequence—M. B. Marcus (p. 179)

The "point of regeneration" method is used to obtain simple sequential equations for determining the complete probability density function for multiple occurrences of events in a Bernoulli sequence. Both the independent and overlapping classes of recurrent events are included in the general framework of these equations. The equations also lead to the generating function for the probability distribution. This is used to obtain the expected recurrent times for the different classes of recurrent events.

A distinction is made between the probability distributions for the occurrence of the kth event at the nth trial and the occurrence of k events in n trials. The latter case is of primary concern.

The methods employed and the results obtained have extensive applications in problems in automatic control, communications, and information processing.

Correspondence (p. 184)

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Annual Index, 1959 (follows p. 191)

Nuclear Science

Vol. NS-6, No. 4, December, 1959

Effects of Reactor Exposure on Boron-Lined and BF$_3$ Proportional Counters—William M. Trenholme (p. 1)

The results of reactor exposure on the characteristics of several types of boron-lined and BF$_3$-filled proportional counters are presented, and the important design and application criteria that lead to improved performance of counters after long reactor exposure are discussed.

The Describing Function of Nuclear Reactors—Henri B. Smets (p. 8)

The integro-differential equation of reactor dynamics is studied and approximated. It is found that a reactor can be represented by means of a linear system and a nonlinear system in cascade. The transfer function of the linear system is the classical low-power transfer function. The nonlinear system is a nonlinear amplifier with an exponential characteristic. The describing function is computed in the case of a water-moderated reactor and it is shown that gain near resonance increases with the magnitude of a reactivity sinusoidal oscillation.

Radiation-Tolerant Electronic Equipment—James R. Burnett (p. 12)

Transistor-Driven Beam Switching Tube Decade Counter—Richard H. Graham (p. 16)

This paper describes an electrical readout decade counter employing the magnetron beam switching tube with transistor drive. Double-pulse resolution is 1 μsec. The unit will accept a variety of transistor types, and will tolerate supply voltage variations of ± 20 per cent at ambient temperatures up to 60°C. A "Pixie" neon indicator is driven without the use of additional transistors. A readout circuit for printer or punched paper tape is presented.

A Stack Effluent Radioisotope Monitor—R. A. Harvey (p. 20)

A system for simultaneously measuring and indicating each of several different radioisotopes in the stack effluent of an atomic energy facility is described. Analog computer techniques are employed for the solution of simultaneous equations and for the counting and scaling. Two types of presentations are included for the output. The total amounts of the radioisotopes emitted to the atmosphere are indicated on registers, and marks are recorded on a moving chart each time specific quantities are emitted. The emission quantity of each isotope is indicated by a register and by a single channel on the chart recorder.

Ultrasonics Engineering

Vol. 7, PGUE-8, February, 1960

Transducer Properties of Lead Titanite Zirconate Ceramics—D. Berlincourt, *et al.* (p. 1)

Physical properties of two commercial ceramics of the lead titanate zirconate family are presented and are compared with data for typical barium titanate ceramics. The importance of dielectric losses is stressed. Preference is given to showing heat generated as a function of nonresonant strain amplitude rather than electric-field amplitude. In this type of presentation, common barium titanate ceramics vary little from each other, while lead titanate zirconate ceramics are quite outstanding.

Thickness-Shear Mode BaTiO$_3$ Ceramic Transducers for Ultrasonic Delay Lines—John E. May, Jr. (p. 7)

The properties of thickness-shear mode BaTiO$_3$ ceramic transducers as applied to ultrasonic delay lines have been investigated. A method of fabricating these transducers has been developed which allows bonding to the delay line by conventional soldering techniques. Compared with thickness-extensional mode transducers, the shear mode transducers have a higher dielectric constant, a lower resistance and a lower capacitance ratio. Substitution for the former on short delay lines at 15 mc increased the bandwidth from 30 per cent to 43 per cent for the same loss. When the quartz transducers on the longer polygon-type lines were replaced with the ceramic transducers, the loss was reduced from 40 db to 20 db.

Vibrations of Ferroelectric Transducer Elements Loaded by Mass and Acoustic Radiation—F. Rosenthal and V. D. Mikuteit (p. 12)

The motion, stress distribution, and electrical impedance are calculated for idealized three-layer sonic transducers, consisting of a piezoelectric element loaded at either end by a mass which may be distributed or lumped, and loaded further by resistive plane-wave acoustic radiation. The piezoelectric element may be driven by use of either the d_{33} or d_{31} effect. The lumped constant equivalent circuit is given for the case of lumped end masses and negligible ceramic mass density. The lumped mass solutions are compared with the somewhat more realistic solution based on the assumption of distributed end masses (obtained on an IBM 650 computer).

Effect of Electrical and Mechanical Terminating Resistances on Loss and Bandwidth According to the Conventional Equivalent Circuit of a Piezoelectric Transducer—R. N. Thurston (p. 16)

This paper includes a rather complete analysis of transducer loss and bandwidth as a function of the mechanical and electrical terminations. For the case of tuning with a shunt coil, Fig. 1 shows lines of constant loss at midband and curves of constant bandwidth plotted with the electrical and mechanical terminations proportional to ordinate and abscissa, respectively. Besides enabling one to find quickly the loss and bandwidth corresponding to any pair of terminations and to visualize the effect of termination changes, this plot also reveals quickly the terminations which have special properties, such as zero loss at midband, zero minimum loss, widest band for a given loss, etc. The case of a series tuning inductor is treated by making use of an approximate correspondence with the shunt case. Results to be expected without a tuning inductor are also indicated.

Ultrasonic Welding Equipment—John N. Antonevich (p. 26)

Little or no work has been reported on investigations into the basic principles of ultrasonic welding. Until these principles are understood, the development of ultrasonic welding equipment will continue to be empirical and the limitations of the process will remain unknown. What is believed to be the first fundamental study of the mechanisms involved in ultrasonic welding has been started. On the basis of observations made in this study, it is postulated that this new welding process is a form of pressure or friction welding, in which material surfaces subjected to a clamping force are welded when made to slide in contact with each other. The process appears to be related directly to the frictional phenomena of galling and seizing.

The relationships between weld strength and the variables of clamping force, duration of weld cycle, ultrasonic motion, and temperature, have been obtained for a Monel sphere welded to a copper plate. This work shows that there is a minimum displacement and temperature required for welding. Also, that weld strength is directly proportional to clamping force, displacement, weld time, and temperature, up to critical values of these variables. Under conditions where critical values are exceeded, weld strength is decreased, apparently because of fatigue about the weld zone. The minimum displacement required for welding agrees with displacements calculated to be necessary to produce sliding.

The fundamental study shows that the essential components of an ultrasonic welder are: 1) an arrangement to clamp together members to be welded, 2) an arrangement to couple vibratory energy to one of the members being welded, 3) a controllable source of vibratory energy, and 4) a timing circuit to control the duration of clamping force and vibratory energy. On the basis of conditions found necessary to produce a weld, the problems encountered in designing welder components are discussed and methods of circumventing them are presented. To improve weldability of materials, welders could be developed to weld materials in vacuo or in an inert atmosphere. This would reduce oxidation around weld zones caused by frictional heating. A control circuit differentiating the temperature at or near the weld site could also be incorporated in ultrasonic welders to control the process and insure uniformity of strength of welds. Present indications are that practical ultrasonic butt, seam, and spot welders can be designed, but they would be limited in application to welding thin-gauge sheets or small components made of ductile materials.

Correction (p. 32)

Contributors (p. 33)

Abstracts and References

Compiled by the Radio Research Organization of the Department of Scientific and Industrial Research, London, England, and Published by Arrangement with that Department and the *Electronic and Radio Engineer*, incorporating *Wireless Engineer*, London, England

NOTE: The Institute of Radio Engineers does not have available copes of the publications mentioned in these papers, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications, not to the IRE.

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The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number. The number in heavy type at the top right is the serial number of the abstract. D.C. numbers marked with a dagger (†) must be regarded as provisional.

U.D.C. NUMBERS

Certain changes and extensions in U.D.C. numbers, as published in PE Notes up to and including PE 666, will be introduced in this and subsequent issues. The changes are:

Artificial satellites:	551.507.362.2	(PE 657)
Semiconductor devices:	621.382	(PE 666)
Velocity-control tubes, klystrons, etc.:	621.385.6	(PE 634)
Quality of received signal, propagation conditions, etc.:	621.391.8	(PE 651)
Color television:	621.397.132	(PE 650)

The "Extensions and Corrections to the U.D.C.," Ser. 3, No. 6, August, 1959, contains details of PE Notes 598–658. This and other U.D.C. publications, including individual PE Notes, are obtainable from The International Federation for Documentation, Willem Witsenplein 6, The Hague, Netherlands, or from The British Standards Institution, 2 Park Street, London, W.1., England.

ACOUSTICS AND AUDIO FREQUENCIES

534.2–8–14 **722**
Propagation of Finite-Amplitude Ultrasonic Waves in a Relaxing Medium—V. A. Krasil'nikov and D. V. Khaminov. (*Akust. Z.* vol. 5, no. 2, pp. 166–169; 1959.) Investigation of ultrasonic propagation at frequencies of 0.5, 1 and 2 mc in acetic acid at various concentrations.

534.2–8–14 **723**
Optical Investigations of the Waveform of Large-Amplitude Ultrasonic Waves in Liquids—V. A. Shutilov. (*Akust. Z.*, vol. 5, no. 2, pp. 231–240; 1959.)

A list of organizations which have available English translations of Russian journals in the electronics and allied fields appears at the end of the Abstracts and References section.

The Index to the Abstracts and References published in the PROC. IRE from February, 1959 through January, 1960 is published by the PROC. IRE, June, 1960, Part II. It is also published by *Electronic Technology* (incorporating *Wireless Engineer*, and *Electronic and Radio Engineer*), and included in the April, 1960 issue of that journal. Included with the Index is a selected list of journals scanned for abstracting with pulishers' addresses.

534.2–8–14 **724**
Propagation of an Acoustic Pulse in an Underwater Sound Channel—N. S. Ageeva (*Akust. Z.*, vol. 5, no. 2, pp. 146–150; 1959.) Investigation of the waveform of 3-ms ultrasonic pulses propagated over distances up to 20 km underwater. Recordings of these pulses are compared with ray diagrams corresponding to the experimental conditions. The pulse shape is observed to vary with the distance from the emitter.

534.213–8 **725**
Velocity and Attenuation of a Narrow-Band, High-Frequency Compressional Pulse in a Solid Waveguide—M. Redwood. (*J. Acoust. Soc. Amer.*, vol. 31, pp. 442–448; April, 1959.) The Pochhammer-Chree solution does not correctly describe the propagation phenomena. A modification of the equations is proposed.

534.22–13 **726**
Velocity of Sound in Relaxing Gases—H. J. Wintle. (*Nature*, vol. 184, suppl. no. 26, pp. 2007–2008; December 26, 1959.) Measurements were made on pure CO_2, O_2 and dry air free from CO_2 of the velocity of sound and corresponding attenuation at frequencies in the range 100–1500 cps at a pressure of one atmosphere and a temperature of 30°C.

534.231 **727**
A New Method of Solving the Problem of a Sound Field in a Fluid Wedge—V. K. Kuznetsov. (*Akust. Z.*, vol. 5, no. 2, pp. 170–175; 1959.) Results are examined for the case of absolutely reflecting boundaries of the wedge. The theory indicates a new phenomenon of normal-wave refraction in the wedge with application to hydroacoustic shore refractions.

534.232–8 **728**
Dependence of the Gain of an Acoustic Focusing System on Sound Intensity—K. A. Naugol'nykh and E. V. Romanenko. (*Akust. Z.*, vol. 5, no. 2, pp. 191–195; 1959.) An expression is obtained for estimating the effects of nonlinear distortion in a focusing system on the amplification coefficient. Miniature receivers are used in investigating the waveform near the focus and the amplification coefficients of three $BaTiO_3$ focusing radiators operating at 0.5, 1.4, and 2.2 mc.

534.26 **729**
Diffraction of a Convergent Cylindrical Wave by a Cylinder—I. N. Kanevskiĭ. (*Akust. Z.*, vol. 5, no. 2, pp. 151–156; 1959.) A mathematical analysis of the diffraction of waves on an infinite cylinder disposed coaxially with the front of the wave. A comparison is made with the case of scatter of a plane wave on a cylinder of small circumference compared to the wavelength.

534.4 **730**
Theory and Practice of Vibration Analysis in the Transient Regime—L. Pimonow. (*Ann. Telecommun.*, vol. 13, pp. 100–111 and 125–152; March/April and May/June, 1958.) An analyzer based on the physical characteristics of the ear is evolved.

534.612:621.352:541.135.6 **731**
Analysis of a Logarithmic Solion Acoustic Pressure Detector—A. F. Wittenborn. (*J. Acoust. Soc. Amer.*, vol. 31, pp. 475–478; April, 1959.) An approximate method is used which avoids solving the Navier-Stokes equation describing the fluid-flow and the convection diffusion equation for calculating the current.

534.75 **732**
Intensive Differential Thresholds for Octave-Band Noise—A. M. Small, Jr, W. E. Bacon, and J. L. Fozard. (*J. Acoust. Soc. Amer.*, vol. 31, pp. 508–510; April, 1959.) Report of an experiment to determine the differential thresholds for octave bands of noise of different center frequencies at various sensation levels.

534.75 **733**
Multiple Observations of Signals in Noise—J. A. Swets, E. F. Shipley, M. J. McKey, and D. M. Green. (*J. Acoust. Soc. Amer.*, vol. 31, pp. 514–521; April, 1959.) Information on the general properties of the hearing process is obtained from experiments on the gain in detectability resulting from repeated observations of a given signal in noise.

534.75:621.391 **734**
Indices of Signal Detectability Obtained with Various Psychophysical Procedures—Swets. (*See* 1013.)

534.76 **735**
Binaural Communication Systems: Preliminary Examination—I. Pollack. (*J. Acoust. Soc. Amer.*, vol. 31, pp. 81–82; January, 1959.) Report of an exploratory study of a binaural system with rejection filtering in which selective operations are transferred to the listener.

534.76 **736**
On the Localization of the Apparent Sound

Source in Two-Channel Stereophonic Transmission—V. S. Man'kovskiĭ. (*Akust. Z.*, vol. 5, no. 2, pp. 176–182; 1959.) Stereophonic tests carried out at frequencies of 200, 1000 and 6000 cps in reverberation rooms of volume 1600 and 1500 meters^3 are described. The dependence of the apparent position of the sound source on the difference in sound level and the coordinates of the listener's position is investigated.

534.78 737

Intelligibility of Reiterated Speech—D. J. Sharf. (*J. Acoust. Soc. Amer.*, vol. 31, pp. 423–427; April, 1959.) An attempt is made to increase the intelligibility of speech sampled in time (interrupted speech) by filling in the blank spaces with repetitions of the speech samples (reiterated speech). There appears to be little advantage in the technique.

534.78:621.391 738

Communication Efficiency of Vocoders—Billings. (*See* 1014.)

534.782 739

An Experimental Transistorized Artificial Larynx—H. L. Barney, F. E. Haworth, and H. K. Dunn. (*Bell. Syst. Tech. J.*, vol. 38, pp. 1337–1356; November, 1959.) A modified telephone receiver driven by a transistor relaxation oscillator, and powered by 10.4-volt Hg cells, is used to produce 0.5-msec pulses at the required pitch. The pitch can be changed and, in use, the telephone receiver is held against the throat.

534.79 740

The Relation between Loudness and the Sound Pressure on the Tympanic Membrane—G. Jahn. (*Hochfrequenz.*, vol. 67, pp. 69–72; November, 1958.) Subjective tests with specially developed equipment show that a given sound pressure at the membrane always results in the same loudness sensation irrespective of the shape of the external sound field.

534.845.1 741

Investigations on the Measurement of the Sound Absorption Coefficient in Echo Chambers—E. Steffen. (*Hochfrequenz.*, vol. 67, pp. 73–77; November, 1958.) The effect of the arrangement and shape of the absorbent material under test on the relation between measured and true absorption coefficients is investigated.

534.861:621.395.665.1 742

Contribution on the Reciprocal Control of Dynamic Range—Völz (*See* 1019.)

534.88 743

On the Detection of Signals in Nonstationary Noise by Product Arrays—J. B. Thomas and T. R. Williams. (*J. Acoust. Soc. Amer.*, vol. 31, pp. 453–462; April, 1959.) Two detection systems having multiple inputs each consisting of nonstationary noise and a stationary random signal are analyzed and compared on the basis of signal/noise ratio.

534.88 744

Correlation of a Finite-Distance Point Source—M. J. Jacobson. (*J. Acoust. Soc. Amer.*, vol. 31, pp. 448–453; April, 1959.) Errors in predicted source directions obtained with a two-receiver steerable correlation system are considered in relation to source distance.

534.88:621.396.965.4 745

Underwater Echo-Ranging with Electronic Sector Scanning: Sea Trials on R.R.S. Discovery II—D. G. Tucker, V. G. Welsby, L. Kay, M. J. Tucker, A. R. Stubbs, and J. G. Henderson. (*J. Brit. IRE*, vol. 19, pp. 681–696; November, 1959.) The general performance of the equipment and its ability to detect fish shoals is described, together with an account of the equipment design. See 3676 of 1958 (Tusher, *et al.*).

621.395.61.089.6:534.6–8 746

Absolute Calibration of Pitot Tubes and Microphones by means of Ultrasonics—V. Gavreau and A. Calaora. (*Ann. Télécommun.*, vol. 13, pp. 153–158; May/June, 1958.) A phase-difference technique for measuring the velocity of air flow is applied to the calibration of pitot tubes at low velocities. The technique is also applied to the absolute calibration of microphones at frequencies below 400 cps.

621.395.623.7:621.318.2 747

Loudspeaker Magnet Design—A. E. Falkus. (*Wireless World*, vol. 66, pp. 41–44; January, 1960.) Magnet design using capped cylindrical slugs of Alcomax III is described.

ANTENNAS AND TRANSMISSION LINES

621.372.2 748

Reflection Coefficients—R. A. Waldron and J. K. Skwirzynski. (*Electronic Radio Eng.*, vol. 36, pp. 464–465; December, 1959.) "The reflection and transmission coefficients are considered for reflecting surfaces which have the same medium on either side. It is shown that the reflection coefficient can be written cos $\theta\cdot e^{j\theta}$ and the transmission coefficient sin $\theta\cdot e^{j(\theta-\pi/2)}$, θ taking values between $-\pi/2$ and $+\pi/2$."

621.372.2.012.11 749

The Smith Chart—R. A. Hickson. (*Wireless World*, vol. 66, pp. 2–9 and pp. 82–85; January and February, 1960.) Derivation and uses of the chart are described.

621.372.2.09 750

Group Velocity and Energy Velocity of Electromagnetic Waves in Loss-Free Lines—W. Kleen and K. Pöschl. (*Arch. elekt. Übertragung*, vol. 12, pp. 567–569; December, 1958.) Energy velocity equals group velocity in two special models of loss-free delay line and in loss-free lines, in general, including those of periodic structure. This is not in agreement with the findings of Borgnis (696 of 1959).

621.372.8:621.77 751

Waveguide Bending Design Analysis—F. J. Fuchs, Jr. (*Bell Syst. Tech. J.*, vol. 38, pp. 1457–1484; November, 1959.) "The art of rectangular tube bending is analyzed, with particular attention being given to tube wall thickness variations. Effects of these variations on tool design are discussed and methods and formulas for determination of wall distortions are presented."

621.372.8.049.7 752

High-Pressure Microwave Window—A. W. Lawson and G. E. Smith. (*Rev. sci. Instr.*, vol. 30, pp. 989–991; November, 1959.) Two conically tapered single crystals of Al_2O_3, with small ends abutting, are used to effect simultaneously a high-pressure seal and a matched transformation from a standard 1-cm circular waveguide at atmospheric pressure to a circular waveguide terminated by a high-Q cavity resonator with an internal hydrostatic pressure up to 10^4 bars.

621.372.81 753

The Impedance Concept in Waveguide Theory—A. Guerbilsky. (*Ann. Télécommun.*, vol. 13, pp. 114–120; May/June, 1958.) Wave propagation studies are simplified by breaking down the propagated wave into uniform plane waves. The method is applied to a parallel-plate and a rectangular waveguide, each containing two dielectrics.

621.372.825 754

Ridge Waveguide Impedance—G. Duckworth. (Proc. IRE, vol. 47, p. 2121; December, 1959.) A simpler formula than that given by Mihran (2425 of 1949) is derived for the characteristic impedance.

621.372.831 755

Mode Conversion—L. Solymar. (*Electronic Radio Eng.*, vol. 36, pp. 461–463; December, 1959.) The electric intensities in the taper section and larger waveguide are determined, and the amplitudes of the spurious modes computed for the case of rectangular waveguides of different cross sections joined by a pyramidal tapered section. A simple formula relating the taper section length to the mode-conversion figure is derived.

621.372.831:621.317.373 756

Phase Measurements through Tapered Junctions—L. Lewin. (*Proc. IEE*, pt. B, vol. 106, pp. 495–496; September, 1959.) The electrical length of a tapered junction matched at the transitions to uniform waveguides is usually taken to be its physical length. A small correction, due to the cylindrical nature of the wave, is calculated theoretically.

621.372.852.22 757

Gyromagnetic Modes in Waveguide Partially Loaded with Ferrite—H. Seidel and R. C. Fletcher. (*Bell Syst. Tech. J.*, vol. 38, pp. 1427–1456; November, 1959.) Analysis is made of all the propagating modes of a vanishingly small rectangular waveguide partially filled with transversely magnetized ferrite. Each of these modes is shown to propagate in only one direction and to tend to be lossy. These properties are of use in the design of a nonresonance isolator.

621.372.855 758

Rectangular-Waveguide Loads—W. G. Voss. (*Electronic Tech.*, vol. 37, pp. 6–8; January, 1960.) Modifications to the construction and mounting of resistive-element dissipative strips when used either as fixed flat loads or as adjustable sliding terminations are described. The two loads are considered to be improvements on any previous design of low-power load, having particular application where a variation of voltage SWR is required over a wide range or where an extremely flat load is required.

621.396.67.095 759

Antenna Power Densities in the Fresnel Region—R. W. Bickmore and R. C. Hansen. (Proc. IRE, vol. 47, pp. 2119–2120; December, 1959.) Curves of maximum power density are presented as a function of distance from the antenna for typical aperture distributions.

621.396.677.43 760

The Horizontal Radiation Patterns of Rhombic Receiving Antennas for Short-Wave Transmission Paths—H. Bohnenstengel, W. Kronjäger, and K. Vogt. (*Nachrichtentech. Z.*, vol. 11, pp. 605–610; December, 1958.) Statistical methods were used to determine the horizontal radiation patterns of a rhombic with 115-meter sides at Eschborn receiving station, for frequencies 5, 10, 15 and 20 mc. Levels of side-lobe and backward radiation were also determined.

621.396.677.71 761

Pattern Synthesis for Slotted-Cylinder Antennas—J. R. Wait and J. Householder. (*J. Res. nat. Bur. Stand.*, vol. 63D, pp. 303–313; November/December, 1959.) Tchebycheff polynomials are used to calculate the distribution of excitation in phase and amplitude of the axial slots, arranged around the circumference of a vertical cylinder, which are necessary to produce the required beamwidth and side-lobe levels in the horizontal plane.

621.396.677.83 762

Passive Microwave Mirrors—R. G. Medhurst. (*Electronic Radio Eng.*, vol. 36, pp. 443–449; December, 1959.) Equations for the phase and amplitude of the near field, close to the main lobe axis, of the primary radiation are derived as modifications of the far-field expressions. Computation of the gain of the antenna-reflector system relative to the antenna alone is simplified; the results agree well with measurements and previously calculated results.

AUTOMATIC COMPUTERS

681.142 763

New Digital-Computer Techniques—(*Proc. IEE*, pt B, vol.106, pp. 444–469; September, 1959.) A report of specialist discussion meetings of the Measurement and Control Section of the IEE, February 16–17, 1959, dealing with character recognition, peripheral equipment, low-temperature storage and switching devices, and special aspects of logical design.

681.142 764

Study of Certain Errors in Analogue Computers—M. W. Overhoff. (*Ann. Télécommun.*, vol. 13, pp. 162–168; July/August, 1958.) Linear and nonlinear imperfections in adding and integrating circuits are related to errors in the solution of differential equations. Analog computers using only ac amplifiers are shown to be a practical proposition.

681.142 765

Electronic "Rotor"—J. S. Johnston. (*Electronic Tech.*, vol. 37, pp. 2–6; January, 1960.) A method is described for introducing the concepts of bearing and angle into fast Cartesian analog computations without the use of mechanical shafts and servomechanisms.

681.142:621.3.029.6 766

Microwave Computer Circuits—F. Leary. (*Electronics*, vol. 32, pp. 77–81; November 20, 1959.) Examples of the use of microwave components and techniques in the construction of simple high-speed computing elements are given. Main problems are the cost and complexity of the waveguide systems required.

681.142:621.376.22 767

Use of a Diode Ring as a Four-Quadrant Multiplier—R. H. Wilcox. *Rev. Sci. Instr.*, vol. 30, pp. 1009–1011; November, 1959.) The operation of a diode ring as a passive analog multiplier is analyzed. Tests show that an accuracy within 1 per cent is obtainable at input levels up to 150 mv.

681.142 768

Electronic Digital Computers [Book Review]—C. V. L. Smith. Publishers: McGraw-Hill, London, 443 pp. (*Electronic Tech.*, vol. 37, pp. 43–44; January, 1960.)

CIRCUITS AND CIRCUIT ELEMENTS

621.316.8:621.391.822 769

Noise of Resistors and Resistor Combinations without and with Load—K. Lunze. (*NachrTech.*, vol. 8, pp. 580–584; December, 1958.) Noise due to current flow in carbon-film-type resistors is particularly considered. The increase over thermal noise is controlled by a factor which depends on loading, film dimensions and the construction of the resistor; from this a formula is derived which contains a constant dependent on film material.

621.319.4 770

The Impedance of Non-quasistationary. Annular-Plate Capacitors—W. Buschbeck (*Telefunken Ztg.* vol. 32, pp. 23–28; March, 1959. English summary, p. 72.) The maximum permissible dimensions for the avoidance of self-resonance are calculated for high-frequency grid stopper and lead-through capacitors.

621.319.42.004.6 771

The Reliability and Life of Impregnated-Paper Capacitors—J. P. Pitts. (*Proc. IEE*, vol. 106, pp. 397–403; October, 1959. Discussion, pp. 403–408.) The construction of paper capacitors and the problems associated with failure in them are reviewed. Possible methods of prolonging life and reducing failures are discussed.

621.372.414 772

Coaxial Resonators with Helical Inner Conductor—W. W. Macalpine and R. O. Schildknecht. (Proc. IRE, vol. 47, pp. 2099–2105; December, 1959.) Resonators of small size can be designed for the VHF and HF ranges. Design formulas are simple, and have been confirmed by experiment.

621.372.44:621.372.6 773

General Energy Relations in Nonlinear Reactances—J. M. Manley and H. E. Rowe. (Proc. IRE, vol. 47, pp. 2115–2116; December, 1959.) An alternative derivation to that used earlier (2988 of 1956) is based on conservation of energy and the properties of the device.

621.372.44:621.372.6 774

A Simplified Derivation of the Manley and Rowe Power Relationships—J. E. Carroll. (*J. Electronics Control*, vol. 6, pp. 359–361; April, 1959.) *See* 773 above.

621.372.5 775

The Determination of Canonical Reactance Quadripoles with Prescribed Iterative Matrix—R. Unbehauen. (*Arch. elekt. Übertragung*, vol. 12, pp. 545–556; December, 1958.) The iterative matrix of any reactance quadripole is realized by means of canonical circuits (see, *e.g.*, 2354 of 1957). The method consists in developing a two-pole function into a continued fraction. This function represents the impedance of the two-pole network which results from terminating by unity resistance the output of a quadripole realizing the given matrix.

621.372.5 776

Applications of Wave Matrices for Calculations on Four-Terminal Networks with Transverse Symmetry—L. R. Yavich. (*Radiotekh. Elektron.*, vol. 4, pp. 341–344; February, 1959.) An extension of the method used in the classical theory of linear networks and based on bisecting the four-terminal network. See 2149 of 1959.

621.372.5:061.3 777

Modern Network Theory in Electrical Engineering—J. T. Allanson. (*Nature*, vol. 184, pp. 1691–1692; November 28, 1959.) Report of a conference held by the Department of Electrical Engineering of the University of Birmingham, September 22–24, 1959.

621.372.5:621.3.016.35 778

Some Considerations on the Stability of Electrical Circuits—I. Gumowski. (*Ann. Télécommun.*, vol. 13, pp. 121–124; May/June, 1958.) A definition of stability based on an integral equation is proposed. It is considered to be more general in applicability than the definitions of Bode and James and Weiss.

621.372.5:621.391.822:621.375.4 779

Contribution to the Theory of Noisy Quadripoles—R. Paul. (*NachrTech.*, vol. 8, pp. 548–568; December, 1958.) Detailed consideration of the parameters which determine the noise figure, particularly of transistors. The effect of circuit configuration on noise figure is discussed and typical circuit parameters for thermal-noise networks are tabulated. The noise characteristics of a single-stage transistor amplifier are investigated.

621.372.5.001.572 780

Transient Synthesizer—P. Eykhoff. (*Electronic Tech.*, vol. 37, pp. 31–36; January, 1960.) An instructional instrument is described which may be used to study the transient response of a quadripole of which the transfer function, in terms of poles and zeros, may be set immediately to the desired form.

621.372.51:621.375.42 781

Power Matching of Quadripoles with Complex Parameters—H. J. Butterweck. (*Arch. elekt. Übertragung*, vol. 12, pp. 540–544; December, 1958.) The passive matching impedances are derived which are required for maximum power amplification with quadripoles having complex parameters, such as transistors at high frequencies. A criterion is established for unconditional stability of a quadripole connected to passive impedances.

621.372.54 782

Monotonicity and Maximally Flat Rational Functions—T. R. O'Meara. (Proc. IRE, vol. 47, pp. 2116–2117; December, 1959.) It is shown that these functions describe the transmission properties for some common cases in loss-free filter design.

621.372.62 783

The Multipole Network with Symmetry of Terminal Number—W. Klein. (*Arch. elekt. Übertragung*, vol. 12, pp. 533–539; December, 1958.) Using matrix notation formulas are derived for crosstalk in multipole networks which have an equal number of input and output terminal pairs.

621.373:621.376.32 784

The Modulation of the Natural Frequency of a Loss-Free Oscillator Circuit—E. Kettel. (*Telefunken Ztg.*, vol. 31, pp. 254–261; December, 1958, and vol. 32, pp. 29–38; March, 1959. English summary.) The modulation can be represented by a Hill's equation. The properties of oscillator circuits with one or two controllable reactance elements are investigated, and equations of the ideal frequency modulator are derived.

621.373.4.029.65:376.23 785

Transistor Phase Detector for Phase-Lock Stabilization of a 30,000-Mc/s Klystron—R. W. Zimmerer. (*Rev. Sci. Instr.*, vol. 30, pp. 1052–1053; November, 1959.) A 15-mc reference signal is applied to the common emitters of two transistors with their bases driven in push-pull by the output of a 15-mc IF amplifier. A control signal is derived from the two collectors and is in series with the klystron repeller voltage supply. See 397 of 1958 (Thompson and Cateora).

621.373.42:621.316.729 786

Oscillator Synchronization at Combination Frequencies for the Purpose of Wide-Band Frequency Stabilization—G. M. Utkin. (*Radiotekh. Elektron.*, vol. 4, pp. 272–285; February, 1959.) Investigation of a system of three coupled oscillators, two of which are tuned to the natural frequency of the third oscillator.

621.373.42:621.316.729 787

Effect of an E.M.F. with Periodically Varying Parameters on an Oscillatory System—N. N. Lunacharskiĭ. (*Radiotekh. Elektron.*, vol. 4, pp. 286–294; February, 1959.) Mathematical analysis, based on the Ricatti equation with variable coefficients, to determine the phase variations in an oscillator with an applied EMF periodically varying in amplitude and frequency.

621.373.421.11 **788**

On the Possibility of Further Generalization in Quasilinear Theory of Single-Circuit LC Oscillators—P. G. Korolev. (*Radiotekh. Elektron.*, vol. 4, pp. 262–271; February, 1959.) A general expression is derived for the complex amplitude of the first harmonic of the oscillator anode current for a given tube characteristic and any arbitrary amplitude values of a synchronous or asynchronous EMF.

621.374.32:621.387.4 **789**

Millimicrosecond Discriminator—D. W. Swift and V. Perez-Mendez. (*Rev. Sci. Instr.*, vol. 30, pp. 1004–1006; November, 1959.) A discriminator circuit for use with counting equipment is described which has a good response to pulses as short as 3 mμsec.

621.374.34 **790**

10^{-9}-sec-Resolving-Time Coincidence Circuit-Limiting Effect in Electron Tubes—F. Lepri. (*Rev. Sci. Instr.*, vol. 30, pp. 1049–1050; November, 1959.) The operation of the circuit described is based on the formation of a virtual cathode between adjacent electrodes under positive grid conditions. Characteristic curves for a pentode Type E 180 F are shown. Tests performed on 50 tubes showed a spread of about 5 per cent in limited-current values.

621.374.4:621.372.44 **791**

Frequency Multiplication with Nonlinear Capacitors—a Circuit Analysis—D. B. Leeson and S. Weinreb. (Proc. IRE, vol. 47, pp. 2076–2084; December, 1959.) From the formulas given, it is possible to specify the optimum nonlinear characteristic for a given circuit and harmonic number, and to calculate the conditions for maximum efficiency. The procedure can be used for frequency division and for circuits with nonlinear inductors.

621.374.4:621.382.22 **792**

Two-Crystal Harmonic Generator—T. E. Hartman. (*Rev. Sci. Instr.*, vol. 30, pp. 1063–1064; November, 1959.) The Si crystals are small polished pieces of microwave-diode material with tungsten-wire cat whiskers. Input and output waveguides have their broad faces in contact. One whole crystal assembly is moved to tune the generator over the fundamental frequency range 22.8–24.3 kmc.

621.375.1.016.35 **793**

Multistage Amplifier Stability Design Criteria—L. G. Cripps. (*Electronic Radio Eng.*, vol. 36, pp. 454–458; December, 1959.) The stability of a cascade amplifier is considered for the two cases: where the stability factors of all the stages are equal a) before cascading, and b) after cascading. A short discussion of the results is given.

621.375.127 **794**

Push-Pull Amplifier Balance—(*Electronic Tech.*, vol. 37, pp. 41–43; January, 1960.) Negative voltage feedback may be applied independently to the two sides of a push-pull amplifier circuit by the insertion of an inductance and shunt resistance network in the lead to the primary center tap of the output transformer.

621.375.22:621.372.44 **795**

Reducing Amplifier Distortion—G. W. Holbrook. (*Electronic Tech.*, vol. 37, pp. 13–20; January, 1960.) A method is described for correcting the nonlinear distortion in tube amplifiers using metallic rectifiers as nonlinear resistances. Examples of corrected single-ended and push-pull amplifiers are given.

621.375.4 **796**

Some Design Considerations for High-Frequency Transistor Amplifiers—D. E. Thomas. (*Bell Syst. Tech. J.*, vol. 38, pp. 1551–1580; November, 1959.) The interaction between output and input due to internal feedback in the transistor is illustrated, and design data are developed for networks used to overcome this. Many practical amplifiers are simple and require a minimum of design effort.

621.375.9:538.569.4 **797**

Spin-Spin Energy Transfer and the Operation of Three-Level Masers—W. B. Mims and J. D. McGee. (Proc. IRE, vol. 47, p. 2120; December, 1959.)

621.375.9:538.569.4 **798**

Construction of a Strong-Field Maser-Type Type Self-Oscillator—C. Fric. (*Compt. rend. Acad. Sci., Paris*, vol. 249, pp. 80–82; July 6, 1959.) Proton spin resonance in a current of magnetically polarized water is applied in the design of a maser which operates in stable conditions without parasitic modulation. Its frequency is 29.66 mc corresponding to a field of 6980 gc. See also 2548 of 1959 (Benoit, *et al.*).

621.375.9:538.569.4 **799**

Operation of a Zero-Field X-Band Maser—J. E. King and R. W. Terhune. (*J. Appl. Phys.*, vol. 30, pp. 1844–1845; November, 1959.) Note on the operation of a 3-level maser amplifier at 12.3 kmc using iron-doped Al_2O_3 and only a small magnetic field for tuning.

621.375.9:538.569.4 **800**

Operating Characteristics of an Ammonia Beam Maser—F. S. Barnes. (Proc. IRE, vol. 47, pp. 2085–2098; December, 1959.) The simple theory is expanded to predict the angular distribution of the molecules leaving the collimator, and the effect of the focuser on the velocity distribution. Comparison with experimental data shows the importance of collisions and beam divergence. 51 references.

621.375.9:621.372.44 **801**

Parametric Amplification and Conversion in Propagating Circuits using Nonlinear Reactances—N. C. Chang. (Proc. IRE, vol. 47, p. 2117; December, 1959.) The case of a time-varying reactance coupling three propagating circuits is analyzed.

621.375.9:621.372.44 **802**

Noise Figure for a Travelling-Wave Parametric Amplifier of the Coupled-Mode Type—C. G. Shafer. (Proc. IRE, vol. 47, pp. 2117–2118; December, 1959.)

621.375.9:621.372.44:621.382.22 **803**

An Extremely Low-Noise 6-kmc/s Parametric Amplifier using Gallium Arsenide Point-Contact Diodes—M. Uenohara and W. M. Sharpless. (Proc. IRE, vol. 47, pp. 2114–2115; December, 1959.) Variation of excess noise temperature with diode temperature has been measured; a range of values comparable to that for a maser has been obtained. *See* also 805 below.

621.375.9:621.372.44:621.382.22 **804**

X-Band Parametric-Amplifier Noise Figures—B. C. De Loach and W. M. Sharpless. (Proc. IRE, vol. 47, p. 2115; December, 1959.) Two types of GaAs point-contact diode with improved gain-bandwidth properties have been examined. *See* also 803 above.

621.375.9:621.372.44:621.382.23 **805**

Low-Noise Parametric Amplifier using Germanium p-n Junction Diode at 6 kMc/s—M. Uenohara and A. E. Bakanowski. (Proc. IRE, vol. 47, pp. 2113–2114; December, 1959.) Experimental arrangements and results are given for noise-figure measurements at various bias voltages and temperatures.

GENERAL PHYSICS

535.62 **806**

Subjective Colour Tests—(*Wireless World*, vol. 66, pp. 33–34; January, 1960.) "Land colour" (3629 of 1959) is found to depend on subjective effects and picture composition to convey the color infomation to the mind of the observer. It does not, therefore, offer a practical system for color television.

537.227:539.2 **807**

Crystal Stability and the Theory of Ferroelectricity—W. Cochran. (*Phys. Rev. Lett.*, vol. 3, pp. 412–414; November 1, 1959.)

537.29:535.34 **808**

Influence of an Electric Field on an Optical Absorption Edge—W. Franz. (*Z. Naturforsch.*, vol. 13a, pp. 484–489; June, 1958.) The absorption edge displacement in a crystal with external field strength of the order of 10^5–10^6 volt/cm is interpreted.

537.311.3:536.2 **809**

Deviation from Matthiessen's Rule and Lattice Thermal Conductivity of Alloys—P. G. Klemens. (*Aust. J. Phys.*, vol. 12, pp. 199–202; June, 1959.) The difference in the ideal electronic thermal conductivity between an alloy and a pure metal can be estimated from the corresponding difference in the ideal electrical resistivity, using the Wiedemann-Franz law.

537.312.62 **810**

The Inverse Skin Effect—M. G. Haines. (*Proc. Phys. Soc.*, vol. 74, pp. 576–584; November 1, 1959.) The axial current density in an infinite cylindrical conductor is calculated as a function of radius and time. Rigid and radially expanding conductors are treated; current density distributions are obtained which sometimes have no resemblance to the normal skin effect.

537.32:536.7 **811**

Irreversible Thermodynamics and Kinetic Theory in the Derivation of Thermoelectric Relations—J. W. Leech. (*Canad. J. Phys.*, vol. 37, pp. 1044–1054; September, 1959.) Considerations of irreversible thermodynamics simplify the determination of thermoelectric coefficients using kinetic-theory arguments based on Boltzmann's equation. A new expression for Thomson's coefficient has been derived. Only simple models have been used but the theory is capable of extension to more complex ones.

537.56 **812**

The Structure of a Stream of Electrons and Ions Drifting and Diffusing in a Gas when Ionization by Collision and Molecular Attachment are Present—L. G. H. Huxley. (*Aust. J. Phys.*, vol. 12, pp. 171–183; June, 1959.)

537.56 **813**

Calculation of the Thermal Conductivity of an Ionic Current—H. Cabannes. (*Compt. rend. Acad. Sci., Paris*, vol. 249, pp. 47–49; July 6, 1959.) A thermal conductivity tensor for the ionic part of a plasma is derived from kinetic theory.

537.56 **814**

Electric Field Distributions in an Ionized Gas—M. Baranger and B. Mozer. (*Phys. Rev.*, vol. 115, pp. 521–525; August 1, 1959.) A method for improving systematically the Holtsmark distribution is described. It is then applied to the calculation of the distribution of the high-frequency component of the electric field in an ionized gas in thermal equilibrium.

537.56:538.56 **815**

Space-Charge Waves in Cylindrical Plasma Columns—A. W. Trivelpiece and R. W. Gould. (*J. Appl. Phys.*, vol. 30, pp. 1784–1793; November, 1959.) The existence of space-charge waves in stationary plasmas of finite cross section has been demonstrated theoretically and experimentally. In addition to forward waves, a plasma can support backward waves when there is a finite axial magnetic field. This makes possible the design of a backward-wave oscillator in which the plasma is the slow-wave circuit. Other possible applications are noted.

537.56:538.63 816

Electromotive Force in a Highly Ionized Plasma Moving Across a Magnetic Field—M. Sakantula, B. E. Clotfelter, W. B. Edwards, and R. G. Fowler. (*J. Appl. Phys.*, vol. 30, pp. 1669–1671; November, 1959.) The EMF is proportional to the flow speed. Oscillographic probe measurements of flow velocity and plasma resistance are described.

538.221 817

Statistical Mechanical Theory of a Random Ferromagnetic System—R. Brout. (*Phys. Rev.*, vol. 115, pp. 824–835; August 15, 1959.) Study of the behavior of solid solutions of paramagnetic impurities which are exchange-coupled in a nonmagnetic substrate yields a considerable amount of information about the nature of the exchange coupling and the temperature dependence of the spin system.

538.222:538.614 818

Study of the Paramagnetic Faraday Effect—J. Soutif-Guicherd. (*Ann. Télécommun.*, vol. 13, pp. 169–185 and 222–238; July/August and September/October, 1958.) From a study of propagation in a gyromagnetic medium the magnitude of the Faraday effect is determined as a function of the constant ξ of the medium. The theory is applied to rotation in paramagnetic salts, and curves are derived for the rotation angle s as a function of the applied field. A description is given of experimental apparatus, and results of measurements on Mn and Fe salts at wavelengths of 1 and 10 cm are compared with with theoretical calculations. 60 references. For an abbreviated report in English see *J. Appl. Phys.*, vol. 29, pp. 256–258; March, 1958.

538.3 819

Structure of the Electromagnetic Field—B. Bertotti. (*Phys. Rev.*, vol. 115, pp. 742–745; August 1, 1959.) A mathematical study of the structure of a curved space-time in which a source-free EM field is present.

538.3:517 820

Helical Fields—H. Poritsky.. (*J. Appl. Phys.*, vol. 30, pp. 1828–1837; November, 1959.) Simple analytic helically invariant solutions of the Laplace equation are given and combined to describe the electrostatic field of a charged helix, and the magnetic field of a helical electric current. Certain graphical flux plotting methods are outlined and illustrated, and network analogies are suggested for solving these fields.

538.3:530.145 821

Significance of Electromagnetic Potentials in the Quantum Theory—Y. Aharonov and D. Bohm. (*Phys. Rev.*, vol. 115, pp. 485–491; August 1, 1959.) Contrary to the conclusions of classical mechanics, there exist effects of potentials on charged particles, even in the region where all the fields vanish. The significance of this is discussed.

538.311 822

Microwave Magnetic Field near a Conducting Perturbation—D. S. Rodbell. (*J. Appl. Phys.*, vol. 30, pp. 1845–1846; November, 1959.) Experiments are described to show that locally intense microwave magnetic-field strengths can be obtained with low power levels by using a conductor to perturb the microwave electric field of a resonant cavity.

538.561:539.12 823

Radiation Produced by the Transit of Charged Particles near Ideally Conducting Bodies—Yu. N. Dnestrovskiĭ and D. P. Kostomarov. (*Radiotekh. Elektron.*, vol. 4, pp. 303–312; February, 1959.) Expressions are obtained for the energy and radiation spectrum. The excitation of oscillations by electrons entering a circular waveguide from an open half-space is discussed.

538.561:539.12:523.165 824

Radiation from Particles Exceeding the Velocity of Light, and Several Applications of it in Experimental Physics—P. A. Čerenkov. [*Science and Culture*, (Calcutta), vol. 25, pp. 281–286; November, 1959.] The history of the discovery and interpretation of Čerenkov radiation is outlined. The application of radiation measurements in the investigation of cosmic-ray showers is discussed.

538.566 825

Vertex-Excited Surface Waves on Both Faces of a Right-Angled Wedge—S. N. Karp and F. C. Karal, Jr. (*Commun. Pure Appl. Math.*, vol. 12, pp. 435–455; August, 1959.)

538.566:535.42]+534.26 826

Diffraction by an Infinite Slit—H. Levine. (*J. Appl. Phys.*, vol. 30, pp. 1673–1682; November, 1959.) A rigorous analysis of the problem for plane waves with special emphasis on short wavelengths and grazing incidence.

538.569.4:621.375.9 827

Construction of a Strong-Field Maser-Type Self-Oscillator—Fric. (*See* 798.)

538.691 828

On the Asymptotic Series Expansion of the Motion of a Charged Particle in Slowly Varying Fields—J. Berkowitz and C. S. Gardner. (*Commun. Pure Appl. Math.*, vol. 12, pp. 501–512; August, 1959.)

539.2:537.122 829

Self-Consistent Field Approach to the Many-Electron Problem—H. Ehrenreich and M. H. Cohen. (*Phys. Rev.*, vol. 115, pp. 786–790; August 15, 1959.) The self-consistent field method in which a many-electron system is described by a time-dependent interaction of a single electron with a self-consistent EM field is shown for many problems to be equivalent to the many-body treatment.

539.2:537.311.1 830

Electrodynamics of Charge Carriers of Negative Effective Mass in Crystals—S Rodriguez. (*Phys. Rev.*, vol. 115, pp. 821–823; August 15, 1959.) The transport properties of negative-effective-mass carriers in crystals are studied. The electrical conductivity of a sample in which the electron distribution function is weakly perturbed from its thermal equilibrium value is always positive, even in the presence of a magnetic field. Cyclotron resonance experiments in equilibrium should therefore display energy absorption.

539.2:538.569.4 831

Optical Pumping in Crystals—H. H. Theissing, P. J. Caplan, F. A. Dieter, and N. Rabbiner. (*Phys. Rev. Lett.*, vol. 3, pp. 460–462; November 15, 1959.) An estimate is made of the expected population change in the ground state of a crystal due to optical pumping.

539.2:538.569.4:621.372.413 832

The Coupling of a Spin System to a Cavity Mode—K. W. H. Stevens and B. Josephson, Jr. (*Proc. Phys. Soc.*, vol. 74, pp. 561–575; November 1, 1959.) A wave-mechanics treatment is given in which the expectation value of the z component of total spin angular momentum is determined. This problem has become important with recent developments in maser technology.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.152.3:621.396.96 833

The Determination of the Electron Density in Interplanetary Space—J. M. Kelso. (*J. Atmos. Terr. Phys.*, vol. 16, pp. 357–359; November, 1959.) The average electron density and the arc length of the ray path from a space vehicle to the ground can be measured by transmitting a pulse to the vehicle on one VHF frequency and receiving pulses emitted from a transponder in the vehicle on this and on a different frequency. In this way, electron densities of the order of 100 electrons/cm^3 can be estimated.

523.16 834

The Temperature required for Nuclear Reactions in Cosmical Electrical Discharges—C. E. R. Bruce. (*Nature*, vol. 184, suppl. no. 26, pp. 2004–2005 December 26, 1959.) It is probable that at a temperature of about 4×10^{80}K nuclear reactions occur freely.

523.164 835

Technical and Astronomical Measurements with the Bonn 25-m Radio Telescope—P. G. Mezger. (*Telefunken Ztg.*, vol. 31, pp. 213–225; December, 1958, and vol. 32, pp. 38–46; March, 1959. English summary.) Details are given of operational tests on the receiving installation of the Bonn University radio telescope on the Stockert [see 3845 of 1957 (Pederzani)]. Additions to the receiving equipment, and the calibration of the antenna by series of astronomical measurements and comparisons with results from other radio telescopes are also described.

523.164 836

The Positional Accuracy of the 25-m Radio Telescope of the Bonn Observatory—B. H. Grahl. (*Telefunken Ztg.*, vol. 31, pp. 226–232; December, 1958. English summary, pp. 272–273.) The tracking system and its adjustment and control are described. Details are also given of the alignment of the telescope axes and the correction of mechanical inaccuracies of the reflector structure.

523.164:551.510.535 837

Radio-Star Scintillations and Ionospheric Disturbances—T. R. Hartz. (*Canad. J. Phys.*, vol. 37. pp. 1137–1152; October, 1959.) Spread-*F* phenomena are not adequate to explain scintillations at 53 mc at Ottawa and precipitation of solar particles seems to be involved at high latitudes.

523.164:551.510.535 838

Cosmic Radio Noise Absorption on 25 Mc/s and F. Scatter—K. R. Ramanathan and R. V. Bhonsle. (*J. Geophys. Res.*, vol. 64, pp. 1635-1637; October, 1959.) A close connection is suggested between the observed anomalous enhancement of scattered VHF signals and the post-sunset increase in the attenuation of cosmic noise.

523.164.3 839

Observation of the Central Region of the Galaxy at 33.3 cm λ with the Large Radio Telescope of the G.A.O. [Principal Astronomical Observatory]—V. G. Malumyan. (*Dokl. Akad. Nauk SSSR*, vol. 129, pp. 1003–1004; December 11, 1959.) A note on radio-telescope observations of Sagittarius made in June, 1959, with a receiver of 60–mc bandwidth and antenna temperature sensitivity <1°K with a 5-second time-constant. Recordings show two brightness components with half-widths respectively 5' and 1.25°. Observations were also made at 3.2 and 9.4 cm λ.

523.164.3 840

Observation of the Sagittarius-A Radio Source at High Resoultion—Yu. N. Pariĭskiĭ.

Dokl. Akad. Nauk SSSR, vol. 129, pp. 1261–1263; December 21, 1959.) Description of observations carried out in April, 1959, by the Pulkovo radio telescope. Graphs show the complex structure of the source. *See* 839 above.

523.164.3 **841**

Neutral Hydrogen Gas in the Taurus-Orion Region observed with a Multichannel 21-cm Line Receiver—R. X. McGee and J. D. Murray. (*Aust. J. Phys.*, vol. 12, pp. 127–133; June, 1959.)

523.164.32:523.165 **842**

Solar Radio Bursts and Low-Energy Cosmic Rays—A. R. Thompson and A. Maxwell. (*Nature*, vol. 185, pp. 89–90; January 9, 1960.) Events following a solar flare are elucidated from a study of the relation between continuum radiation [1126 *1* of 1958 (Maxwell, *et al.*)] and the occurrence of enhanced proton flux at the earth.

523.164.32:523.755 **843**

On the Correlation of Solar Noise Fluctuations in Harmonically Related Bands—L. R. O. Storey. (*J. Res. Nat. Bur. Stand.*, vol. 63D, pp. 293–296; November/December, 1959.) A method is proposed for the study of the solar corona; a delayed correlation between rapid fluctuations of enhanced solar radio emission (Type II) in harmonically related frequency bands would be observed.

523.164.4 **844**

Occultations of the Crab Nebula by the Solar Corona in June 1957 and 1958—O. B. Slee, (*Aust. J. Phys.*, vol. 12, pp. 134–156; June, 1959.) The distribution of 85.5-mc radiation on days when the angular separation is less than 10 solar radii is not consistent with a symmetrical scattering process; better agreement is obtained by postulating the existence of scattering and regular refraction of comparable magnitude.

523.165 **845**

Energy Spectrum of the Heavy Nuclei in the Cosmic Radiation between 7- and 100-BeV/Nucleon—P. L. Jain, E. Lohrmann, and M. W. Teucher. (*Phys. Rev.*, vol. 115, pp. 654–659; August 1, 1959.) From observations it is concluded that a power spectrum of the form $E^{1 \cdot 6}$ (where E = energy/nucleon) holds for heavy nuclei energies up to 100 bev/nucleon and for α particles up to 7 bev. With certain assumptions, the same exponent is a satisfactory value for the spectrum of α particles for energies up to 1500 bev.

523.165 **846**

The Onset Times of Forbush-Type Cosmic-Ray Intensity Decreases—A. G. Fenton, K. G. McCracken, D. C. Rose, and B. G. Wilson. (*Canad. J. Phys.*, vol. 37, pp. 970–982; September, 1959.)

523.165 **847**

A Comparison of the Cosmic-Ray Intensity at High Altitudes with the Nucleonic Component at Ground Elevation—J. E. Henkel, J. A. Lockwood, and J. H. Trainor. (*J. Geophys. Res.*, vol. 64, pp. 1427–1438; October, 1959.) Results of balloon-borne soundings with single Geiger tubes are compared with those made by a nucleonic detector on Mount Washington at 1909-meter height.

523.165 **848**

On the Structure of Extensive Cosmic-Ray Air Showers: the Penetrating Component—E. W. Kellermann and N. Dickinson. (*Proc. Phys. Soc.*, vol. 74, pp. 554–560; November 1, 1959.)

523.165 **849**

Momentum Spectrum of the Van Allen Radiation—H. Alfvén. (*Phys. Rev. Lett.*, vol. 3, pp. 459–460; November 15, 1959.) The spectrum appears to obey the same power law as cosmic radiation indicating that the Van Allen radiation may be explained by the mechanism responsible for the acceleration of cosmic radiation.

523.3 **850**

Absence of Craters on the Far Side of the Moon—D. B. Beard. (*Nature*, vol. 184, suppl. no. 21, p. 1631; November 21, 1959.) The reported lower incidence of craters on the far side of the moon may be explained by the effect of the earth's gravitational field on meteoric material in orbit about the sun.

523.3:551.507.362.1 **851**

Observations related to the Impact of Lunik II—G. Fielder. (*Nature*, vol. 185, p. 11; January 2, 1960.) A table is given of the results of a number of independent observations of the impact of Lunik II with the moon. See also 4051 of 1959 (Davies and Lovell).

523.:551.507.362.2 **852**

First Results Derived from the Photographs of the Moon's Far Side—N. P. Barabashov and Yu. N. Lipskiĭ. (*Dokl. Akad. Nauk SSSR*, vol. 129, pp. 1000–1002; December 11, 1959.) Photographs taken by the third soviet cosmic rocket launched October 4, 1959, are shown. During 40 minutes of camera operation, numerous photographs were taken of the moon's far side at distances 65,200 to 68,400 km from its surface.

523.5:621.396.9 **853**

The Temporal Variation of the Heights of Reflection Points of Meteor Trails—A. A. Weiss. (*Aust. J. Phys.*, vol. 12, pp. 116–126; June, 1959.) There is no significant seasonal variation in either the mean height or the width of the height distribution. The mean height alone shows a diurnal variation with maximum height near midnight. Both height parameters depend on the zenith angle of the apex of the earth's way.

523.5:621.396.9 **854**

Air Motions and the Fading, Diversity, and Aspect Sensitivity of Meteoric Echoes—L. A. Manning. (*J. Geophys. Res.*, vol. 64, pp. 1415–1425; October, 1959.) A theory is given of meteoric reflection from distorted trails. Experimental results confirm the predictions and facilitate a determination of the principal properties of the wind profile.

523.5:621.396.9 **855**

Meteor Radiation Distributions and the Radio-Echo Rates Observed by Forward Scatter—M. L. Meeks and J. C. James. (*J. Atmos. Terr. Phys.*, vol. 16, pp. 228–235; November, 1959.) The diurnal variations observed by forward scatter over three different paths are found to be in reasonable agreement with predictions made using a simplified model distribution based on the radar data of Hawkins (*Astrophys. J.*,vol. 61, p. 386; 1956.

523.5:621.396.9 **856**

Turbulence at Altitudes of 80–100 km and its Effects on Long-Duration Meteor Echoes—J. S. Greenhow and E. L. Neufeld. (*J. Atmos. Terr. Phys.*, vol. 16, pp. 384–392; November, 1959.) The characteristics of the echoes can be readily explained in terms of multiple reflections from overdense columns of ionization and are not consistent with the theory of incoherent scattering from underdense columns.

523.746.5 **857**

Prediction of Sunspot Numbers for Cycle 20—W. B. Chadwick. (*Nature*, vol. 184, suppl. no. 23, p. 1787; December 5, 1959.) Comment on 3282 of 1959 (Herrinck).

550.38 **858**

Note on Conjugate Points of Geomagnetic Field Lines for some Selected Auroral and Whistler Stations of the I. G. Y.—E. H. Vestine. (*J. Geophys. Res.*, vol. 64, pp. 1411–1414; October, 1959.) A method of deriving conjugate points of some auroral and magnetic stations of the I. G. Y.

550.385 **859**

Geomagnetic Oscillations at Middle Latitudes: Parts 1 and 2—E. Maple (*J. Geophys. Res.*, vol. 64, pp. 1395–1409; October, 1959.) Observations of geomagnetic fluctuations in the frequency range 1–0.005 cps, are discussed, based on measurements at Tucson,Ariz. The results favor the hypothesis of intralayer hydromagnetic resonance in the ionosphere as the source mechanism.

550.385 **860**

Geomagnetic Activity at Halley Bay on Disturbed Days—J. MacDowall.(*Nature*, vol. 184, pp. 1543–1545; November 14, 1959.) Diurnal variations of geomagnetic activity on disturbed days are related to auroral and ionospheric effects. An analysis of five worldwide sudden storm commencements (SSC) shows positive departures of the horizontal field before midnight and negative departures after midnight.

550.385:550.37 **861**

The Relation between *H*- and *Z*-Variations near the Equatorial Electrojet—C. A. Onwumechilli. (*J. Atmos. Terr. Phys.*, vol. 16, pp. 274–282; November, 1959.) Small fluctuations in H and Z are strongly correlated particularly during daylight hours. The ratio of the amplitudes of the components during these fluctuations shows a marked dirunal variation but is almost constant during the day.

550.385.26 **862**

Note on the Tidal Theory of the S_q Magnetic Field—R. L. Ingraham. (*J. Atmos. Terr. Phys.*, vol. 16, pp. 263–273; November, 1959.) The solar quiet day magnetic field has a diurnal component which is roughly a thousand times greater than expected on tidal theory. Possible explanations are given.

550.385.4 **863**

Hydromagnetic Propagation of Sudden Commencements of Magnetic Storms—W. E. Francis, M. I. Green, and A. J. Dessler. (*J. Geophys. Res.*, vol. 64, pp. 1643–1645; October, 1959.) A brief analysis of hydromagnetic propagation, which gives an expression for the transit time of waves traveling in the earth's equatorial plane.

550.389.2 **864**

The International Geophysical Year 1957–58—D. Brunt. (*Proc. IEE*, pt. B, vol. 106, pp. 437–443; September, 1959.) Research carried out during the I.G.Y. in different branches of geophysics is reviewed. Attention is drawn to new concepts already established and to those probably true.

551.507.362.2+629.19 **865**

Space Probes and Satellites—[*Engineering*, (*London*), vol. 187, pp. 615–616; May 8, 1959.] A summary of the main details of all satellite and space-vehicle launchings up to April 13, 1959.

551.507.362.2 **866**

Temperature Stabilization of Highly Reflecting Spherical Satellites—G. Hass, L. F. Drummeter, Jr., and M. Schach. (*J. Opt. Soc. Amer.*, vol. 49, pp. 918–924; September, 1959.)

551.507.362.2 **867**

A Correction Necessary for the Application of the Doppler Effect to the Measurements of

Distances to Satellites—J. M. Brito, Jr. (Proc. IRE, vol. 47, p. 2023; November, 1959.) The use of spherical-earth geometry gives a value for minimum passing range considerably lower than that obtained from plane-earth geometry.

551.507.362.2 868

The Fluctuations of the Acceleration of Satellites—H. K. Paetzold. (*J. Atmos. Terr. Phys.*, vol. 16, pp. 259–262; November, 1959.) The fluctuations are associated with changes in air density at perigee height. These density variations may be associated with changes in sunspot activity; they are, in any case, certainly due to variable heating at the perigee height.

551.507.362.2 869

A Method for Measuring Local Electron Density from an Artificial Satellite—L. R. O. Storey. (*J. Res. Nat. Bur. Stand.*, vol. 63D, pp. 325–340; November /December, 1959.) A proposal for a satellite equipped to measure the electric and magnetic fields in the radio wave set up in the neighborhood of the satellite by "whistler"-mode transmission from a ground-based VLF transmitter. From the wave impedance given by the ratio of the fields the local electron density could be deduced.

551.510.52 870

Applications of the Molecular Refractivity in Radio Meteorology—B. R. Bean and R. M. Gallet. (*J. Geophys. Res.*, vol. 64, pp. 1439–1444; October, 1959.) A discussion of the advantages of molecular refractivity as a significant parameter in studies of climatic differences and air mass characteristics.

551.510.52:621.391.812.621.1 871

Radio-Refractive-Index Climate Near the Ground—B. R. Bean and J. D. Horn. (*J. Res. Nat. Bur. Stand.*, vol. 63D, pp. 259–271; November/December, 1959.) World-wide maps of refractive index contours are given. The index varies systematically with climate, and is most accurately deduced from charts of the reduced-to-sea-level index.

551.510.52:621.391.812.621.1 872

Central Radio Propagation Laboratory Exponential Reference Atmosphere—B. R. Bean and G. D. Thayer. (*J. Res. Nat. Bur. Stand.*, vol. 63D, pp. 315–317; November/December, 1959.) A model for the variation of atmospheric refractive index with height is proposed which gives more accurate radio-ray profiles than the 4/3 earth radius treatment. A sample table of ray profiles is given.

551.510.53 873

Analytic and Experimental Electrical Conductivity between the Stratosphere and the Ionosphere—R. E. Bourdeau, E. C. Whipple, Jr, and J. F. Clark. (*J. Geophys. Res.*, vol. 64, pp. 1363–1370; October, 1959.) Results for the height range 35–80 km obtained from rocket soundings are compared with predicted values based on ion equilibrium and ionization by cosmic rays only.

551.510.535 874

Detection of an Electrical Current in the Ionosphere above Greenland—L. J. Cahill, Jr. (*J. Geophys. Res.*, vol. 64, pp. 1377–1380; October, 1959.) A description of measurements made on August 6, 1957, using a rocket-borne magnetometer. The average current density of the layer detected was about 11A/km².

551.510.535 875

Langmuir Probe Measurements in the Ionosphere—R. L. Boggess, L. H. Brace, and N. W. Spencer. (*J. Geophys. Res.*, vol. 64, pp. 1627–1630; October, 1959.) Measurements are described of electron temperature and positive-ion number density of the *E* layer above Fort Churchill, Manitoba, Canada, on November 30, 1958.

551.510.535 876

Note on Quiet-Day Vertical Cross-Sections of the Ionosphere along 75°W Geographic Meridian—J. W. Wright. (*J. Geophys. Res.*, vol. 64, pp. 1631–1634; October, 1959.) A discussion of results now being obtained from stations on the 75th meridian.

551.510.535 877

Gyro-Frequency in the Ionospheric Regions—S. Datta and R. N. Datta. (*Indian J. Phys.*, vol. 33, pp. 316–324; July, 1959.) Experimental results show that the magnetic fields calculated from gyro frequencies in the *E*, *F*, and F_2 regions are higher than expected. The *E*-region gyrofrequency has a marked semidiurnal variation with a minimum at midday.

551.510.535 878

An Investigation of the Ionospheric D Region—J. A. Fejer and R. W. Vice. (*J. Atmos. Terr. Phys.*, vol. 16, pp. 291–306; November, 1959.) A model of the lower ionosphere between 68 km and 85 km is obtained from the observation of weak echoes from the D region and the measurement of ionospheric wave interaction.

551.510.535 879

Some Measurements of Collision Frequency in the E Region of the Ionosphere—D. M. Schlapp. (*J. Atmos. Terr. Phys.*, vol. 16, pp. 340–343; November, 1959.) Measurements are made by observing how the deviative absorption of a radio echo varies with changes in its group path as a critical frequency is approached. Collision frequency at a fixed height seems to increase with solar activity.

551.510.535 880

E-Region Winds—W. H. Ward. (*J. Atmos. Terr. Phys.*, vol. 16, pp. 394–395; November 1959.) The suggestion that heating of the E, region by the sun is responsible for E-region winds is supported by some experimental results. See also 3259 of 1955 (Greenhow and Neufeld).

551.510.535 881

A Relation between Giant Travelling Disturbances and Sporadic-E Ionization—L. H. Heisler. (*Nature*, vol. 184, suppl. no. 23, pp. 1788–1789; December 5, 1959.) Ionosonde records for the period July 10–28, 1955, show a relation between E_s anomalies and traveling disturbances. Sequential E_s [1899 of 1951 (McNicol and Gipps)] and F_2 cusp-type anomalies [1434 of 1957 (Munro and Heisler)] are also discussed in relation to these disturbances.

551.510.535 882

Two Anomalies in the Behaviour of the F_2 Layer of the Ionosphere—S. Croom, A. Robbins, and J. O. Thomas. (*Nature*, vol. 184, suppl. no. 26, pp. 2003–2004; December 26, 1959.) Report of a study of electron-density data at a number of widely distributed observatories during equinox in a year of high sunspot activity. Results show that the "geomagnetic anomaly" described by Appleton (1337 of 1955) extends much farther from the magnetic equator as lower heights are investigated. A diurnal asymmetry also exists, with forenoon electron densities greater than corresponding afternoon ones.

551.510.535 883

Some Remarks on Motion of Ionospheric Irregularity—N. Matuura. (*Rept. Ionosphere Space Research Japan*, vol. 13, pp. 4–20; March, 1959. Discussion.) The relative motion between irregularities in electron density and their surroundings is discussed. Two treatments of the problem are considered: the surrounding ionization being assumed to be a) a hydromagnetic fluid, and b) a fully ionized gas.

551.510.535 884

The Motion of Irregularities in the Ionosphere—T. Tsuda. (*Rept. Ionosphere Space Research Japan*, vol. 13, pp. 56–61; March, 1959. Discussion.) The mean velocity of irregularities in electron density is determined using simple turbulence theory. The irregularities move with the same velocity as the winds in the E region but there is a marked difference from wind velocity in the F region.

551.510.535 885

World-Wide Measurements of Horizontal Ionospheric Drifts—T. Shimazaki. (*Rept. Ionosphere Space Research Japan*, vol. 13, pp. 21–47; March, 1959. Discussion.) A critical survey is made of radio techniques for measuring horizontal drifts. A comparison is then made of results obtained during the I.G.Y. using the closely-spaced-receiver method which discloses seasonal and latitudinal differences in the diurnal variation of drifts. Tentative explanations of these results are given.

551.510.535 886

Ionospheric Drifts at Yamagawa in Japan—K. Tsukamoto and Y. Ogata. (*Rept. Ionosphere Space Research Japan*, vol. 13, pp. 48–55; March, 1959. Discussion.) Drift speeds, diurnal variations of the drift vector, and the relation between speed and geomagnetic activity are discussed using results obtained during about one year from August, 1957.

551.510.535 887

Ionized Irregularities and Wind Motion in the Ionosphere—S. Kato. (*Rept. Ionosphere Space Research Japan*, vol. 13, pp. 62–78; March, 1959. Discussion.) The motion of irregularities in electron density are studied on the basis of a three-dimensional model. Electrodynamic effects are slight in the E region under magnetically quiet conditions but are appreciable in the F region and even in the E region for disturbed conditions at high latitudes. A general discussion on the stability and lifetime of regions of irregular electron density is given.

551.510.535 888

Horizontal Winds and Ionization Drifts in the Ionosphere—H. Maeda. (*Rept. Ionosphere Space Research Japan*, vol. 13, pp. 79–90; March, 1959. Discussion.) From a study of geomagnetic observations during the International Polar Year, 1932–1933, it is concluded that ionization drifts in the F region are due to an applied electric field and not to air movement, while in the E region the ionization drift is almost the same as the air movement.

551.510.535 889

Horizontal Drifts in the E Region at Waltair—R. R. Rao and B. R. Rao. (*Nature*, vol. 185, pp. 27–28; January 2, 1960.) Results of an analysis of data obtained over a two-year period. A 24-hourly east-west component predominates.

551.510.535 890

Study of "Winds" in the F Region of the Ionosphere during the Unusual Days in the I.G.Y. Calendar—R. N. Singh and S. R. Khastgir. (*J. Atmos. Terr. Phys.*, vol. 16, pp. 376–383; November, 1959.) Fading records were taken at 3.8–4.2 mc on three spaced receivers at Banaras. The experimental results are described and the fading records discussed with reference to the random and steady movement of irregularities.

551.510.535:523.5:621.396.96 891

A Theory for determining Upper-Atmos-

phere Winds from Radio Observations on Meteor Trails—G. V. Groves. (*J. Atmos. Terr. Phys.*, vol. 16, pp. 344–356; November, 1959.)

551.510.535:523.78 892

The Effects of a Solar Eclipse on a Stratified Ionosphere—J. A. Gledhill. (*J. Atmos. Terr. Phys.*, vol. 16, pp. 360–366; November, 1959.) The form of the isoelectronic surfaces in a three-layer ionosphere during an eclipse have been calculated. Tilts in these surfaces may lead to difficulties in the interpretation of ionograms.

551.510.535:523.78 893

The Behaviour of the Ionosphere over Cape Town and Johannesburg during the Annular Solar Eclipse of 25 December 1954—J. A. Gledhill. (*J. Atmos. Terr. Phys.*, vol. 16, pp. 367–375; November, 1959.) It was not possible to explain the ionospheric changes at Cape Town, Johannesburg and Grahamstown in terms of a single solar brightness model. Oblique reflections may have affected the ionograms. There is some evidence that $\alpha'E = 4 \times 10^{-8} \text{cm}^3 \text{second}^{-1}$.

551.510.535:621.391.812.63 894

The Frequency Dependence of Ionospheric Absorption—K. Bibl, A. Paul, and K. Rawer. (*J. Atmos. Terr. Phys.*, vol. 16, pp. 324–339; November, 1959. In German.) The total absorption was formerly expressed as a function of nondeviative D-layer plus deviative E-layer absorption [985 of 1952 (Bibl and Rawer)]. The expression has been modified to include the effect of a thin E_s layer. Experimental data are discussed in the light of the new theory.

551.510.535:621.391.812.63 895

The Absorption of Short Radio Waves in the D, E and F Regions of the Ionosphere—J. D. Whitehead. (*J. Atmos. Terr. Phys.*, vol. 16, pp. 283–290; November, 1959.) Values of noon absorption at Slough from 1947 to 1953 have been analyzed. The electron collision frequencies for the E and F regions have been calculated and the variations of D-region absorption with sunspot number examined. The significance of the results is discussed.

551.510.535:621.391.812.63 896

The Use of Full-Wave Solutions in the Interpretation of Ionospheric Absorption Measurements—J. A. Fejer and R. W. Vice. (*J. Atmos. Terr. Phys.*, vol. 16, pp. 307–317; November, 1959.) A full-wave correction is given which can be made to the usual Appleton-Hartree absorption calculation. Examples show that for a frequency of 2 mc this correction cannot be neglected.

551.510.535:621.396.96:523.3 897

Measurements of Ionospheric Electron Content by the Lunar Radio Technique—S. J. Bauer and F. B. Daniels. (*J. Geophys. Res.*, vol. 64, pp. 1371–1376; October, 1959.) Measurements of the Faraday rotation of lunar radio echoes at 151 mc are used to determine the time variation in the total ionospheric electron content. Absolute values of electron content are derived from data on the electron content below the F_2 peak computed from vertical-incidence experiments.

551.593 898

Excitation Mechanisms of the Oxygen 5577 Emission in the Upper Atmosphere—E. Tandberg-Hanssen and F. E. Roach. (*J. Res. Nat. Bur. Stand.*, vol. 63D, pp. 319–324; November/December, 1959.) Photochemical reactions affected by mass motions, and direct excitation due to mass motions are examined.

551.594.5+551.593 899

Auroral and Nightglow Observations at Ås, Norway—G. Kvifte. (*J. Atmos. Terr. Phys.*, vol. 16, pp. 252–258; November, 1959.) More precise wavelength measurements are given for bands of the Meinel OH system in the nightglow.

551.594.5 900

Photometric Observations of Subvisual Red Auroral Arcs at Middle Latitudes—R. A. Duncan. (*Aust. J. Phys.*, vol. 12, pp. 197–198; June, 1959.) The arcs have been observed on the night following a bright visual aurora at a time when the magnetic disturbance index had fallen to five or less.

551.594.5 901

The Southern Auroral Zone in Geomagnetic Longitude Sector 20°E—S. Evans and G. M. Thomas. (*J. Geophys. Res.*, vol. 64, pp. 1381–1388; October. 1959.) Visual auroral observations at Antartic bases are tabulated to show the variation of the frequency of aurora occurrences with geomagnetic latitude between 63° and 79°.

551.594.5 902

Antartic Auroral Observations, Ellsworth Station, 1957—J. M. Malville. (*J. Geophys. Res.*, vol. 64, pp. 1389–1393; October, 1959.) A summary of auroral observations made during the antarctic winter of 1957. Evidence is obtained of a spiraling zone of maximum auroral activity.

551.594.5 903

Spectroscopic Observations of the Great Aurora of 10 February 1958: Parts 1 and 2—K. C. Clark and A. E. Belon. (*J. Atmos. Terr. Phys.*, vol. 16, pp. 205–227; November, 1959.) Spectrograms show an enhancement of vibrational excitation in the $B(F^2\Sigma_u^+)$ state of N_2^+ which can be attributed to a charge exchange process in a strong proton shower. These spectrograms also show a large number of atomic lines seldom seen, most of which are due to O*I* and N *II* transitions.

551.594.5 904

A Monochromatic Low-Latitude Aurora—F. E. Roach and E. Marovich. (*J. Res. Nat. Bur. Stand.*, vol. 63D, pp. 297–301; November /December, 1959.) A particular monochromatic auroral arc (6300Å) which occurred over Colorado, September 29–30, 1957, is described. It seems to have been a continuation of a similar one observed in France.

551.594.6 905

Telluric Origin of the Whistler Medium—F. S. Johnson. (*Nature*, vol. 184, suppl. no. 23, pp. 1787–1788; December 5, 1959.) Arguments are presented to show that the medium cannot be of solar or interplanetary origin and a mechanism is suggested for a telluric origin.

551.594.6 906

Effect of Latitude on the Diurnal Maximum of "Dawn Chorus"—J. H. Pope. (*Nature*, vol. 185, pp. 87–88; January 9, 1960.) Observations of the local time of diurnal maximum have been made at 16 stations. Analysis based on an eccentric dipole field is useful when considering "dawn chorus" theories.

521.030:537.5 907

Cosmic Electrodynamics. (Book Review)—J. W. Dungey, Publishers: University Press, Cambridge, 183 pp.; 1958. (*Nature*, vol. 184, pp. 1672–1673; November 28, 1959.) This book, bearing a similar title to that by Alfvén (see 2777 of 1950), indicates the progress and consolidation achieved in the field during the intervening decade.

LOCATION AND AIDS TO NAVIGATION

621.396.96 908

Optical Simulation of Radar Resolution—R. D. Rawcliffe, W. W. Lichtenberger, and H. V. Krone. (*J. Opt. Soc. Amer.*, vol. 49, pp. 887–890; September, 1959.) An optical simulator has been constructed in which the high resolution of an aerial photograph is degraded by known amounts in order to determine the radar resolution required to achieve a given target detectability.

621.396.96:621.396.934 909

Low-Cost Active Radar for Miss-Distance Data—W. H. Doty. (*Electronics*, vol. 32, pp. 91–93; November 20, 1959.) Frequency-sweep radar located within the target gives a direct indication of the miss distance.

621.396.965.4:534.88 910

Underwater Echo-Ranging with Electronic Sector Scanning: Sea Trials on *R.R.S. Discovery II*—Tucker, Welsby, Kay, Tucker, Stubbs, and Henderson. (*See* 745.)

621.396.967 911

The Line Elements between Transmitter and Aerial of a Radar Installation and their Influence on Transmitter Frequency Stability—H. Schwindling. (*Telefunken Ztg*, vol. 31, pp. 242–253; December, 1958. English summary, pp. 273–274.) An airport surveillance radar installation is examined with regard to stability conditions governed by the "long-line" effect (3653 of 1958). The maintenance of stability by means of phase shifters can only be carried out by referring to the frequency spectrum of the transmitted pulses.

MATERIALS AND SUBSIDIARY TECHNIQUES

535.215 912

The Theory of Glow Curves—K. W. Böer, S. Oberländer, and J. Voight. (*Z. Naturforsch.*, vol. 13a, pp. 544–547; July, 1958.) The analysis of trapping-center distribution on the basis of glow curves and allowing for retrapping is considered. See also 3321 of 1959.

535.215 913

Gain-Bandwidth Product of Photoconductors—R. W. Redington. (*Phys. Rev.*, vol. 115, pp. 894–896; August 15, 1959.) Extension of earlier treatment (1914 of 1958) to include an arbitrary distribution of impurity levels. Results show the reciprocal of the relaxation time to be an upper limit for the gain-bandwidth product of a photoconductor with space-charge-limited dark current.

535.215:546.48'221 914

Dislocations in CdS Crystals and their Significance in Photoconductivity—E. Votava. (*Z. Naturforsch.*, vol. 13a, pp. 542–544; July, 1958. plate.) Methods of growing crystals with pronounced dislocations and of making these visible are discussed. Experiments seem to confirm that dislocations may reduce the electron lifetime.

535.215:546.48'221 915

The Electrical Behaviour of CdS Single Crystals with Voltage Pulses in the Breakdown Region—K. W. Böer, J. Dziesiaty, and U. Kümmel. (*Z. Naturforsch.*, vol. 13a, pp. 560–562; July, 1958.) Crystal current is plotted as a function of pulse width, amplitude and repetition frequency and comparisons are made with dc measurements at breakdown field strength. [see also 2749 of 1958 (Böer and Kümmel)].

535.215:546.863'221 916

Cathodoconductivity of Antimony Sulphide—Ya. A. Oksman and G. P. Tikhomirov. (*Radiotekh. Elektron.*, vol. 4, pp. 344–346; February, 1959.) Investigation of the possibility of increasing the conductivity of thin dielectric and semiconductor films by electron bombardment for the purpose of amplifying the photoelectron emission current. The limiting values of the gains which can be obtained using these films at room temperature are determined. See 2458 of 1957 (Oksman).

535.37:546.47'221 917

Optical Properties of Activated and Unactivated Hexagonal ZnS Single Crystals—S. P. Keller and G. D. Pettit. (*Phys. Rev.*, vol. 115, pp. 526–536; August 1, 1959.) Further measurements on the polarization effects in the transmission spectrum of unactivated ZnS show that theoretical prediction holds except in the region 290–325 mμ λ. The wavelength dependence and the polarization of the excitation and fluorescence spectra of the activated and unactivated crystals were measured at room temperature and at 77°K.

535.376 918

Energy Limit for Electroenhancement and Electroextinction Effects—G. Destriau. (*Compt. rend. Acad. Sci., Paris*, vol. 249, pp. 245–247; July 15, 1959.) The sensitivity of luminescent compounds of the CdZnS-Mn, *X* type under electron bombardment is enhanced by X-ray and reduced by ultraviolet-ray excitation. Extrapolation of experimental curves indicates that with an activation energy below 5000 ev no electroenhancement occurs.

535.376 919

New Effect of Sensitization to the Action of X Rays, by means of Infrared, in Luminescent Products of the Type CdZnS-Mn, *X*—F. Pingault. (*Compt. rend. Acad. Sci., Paris*, vol. 249, pp. 248–250; July 15, 1959.) Infrared radiation produces an enhancement effect similar to that described by Destriau (*see, e.g.* 918 above). The analogy between the action of an applied electric field and infrared irradiation is stressed.

535.376:546.48'221 920

Some Effects of Low Fields on Luminescence of CdS—C. E. Bleil and D. D. Snyder. (*J. Appl. Phys.*, vol. 30, pp. 1699–1702; November, 1959.) Experimental results are given for the shift of cathodoluminescence peaks with applied field and for the dependence of conductivity on voltage and irradiation. An explanation is suggested.

537.227 921

Asymmetric Hysteresis Loops and the Pyroelectric Effect in Triglycine Sulphate—A. Savage and R. C. Miller. (*J. Appl. Phys.*, vol. 30, pp. 1646–1648; November, 1959.) The apparent asymmetry, as disclosed by pyroelectric measurements, is shown to be due to electrode edge effects.

537.227 922

Domain Processes in Lead Titanate Zirconate and Barium Titanate Ceramics—D. Berlincourt and H. H. A. Krueger. (*J. Appl. Phys.*, vol. 30, pp. 1804–1810; November, 1959.) An experimental investigation of domain reorientation during poling, and domain switching processes under high mechanical or electric stress.

537.227:546.431'824–31 923

On the Dependence of the Switching Time of Barium Titanate Crystals on their Thickness—M. E. Drougard and R. Landauer. (*J. Appl. Phys.*, vol. 30, pp. 1663–1668; November, 1959.) Analysis of domain motion and field calculations show that the dependence of switching rate on thickness is explained by a monomolecular surface layer having an irreversible polarization and a low dielectric constant. Implications of this in explaining the switching mechanism are discussed.

537.311.31:538.63 924

Magnetoresistance of Copper—J. de Launay, R. L. Dolecek, and R. T. Webber. (*J. Phys. Chem. Solids*, vol. 11, pp. 37–42; September, 1959.) Report and discussion of measurements at 297°, 78° and 4.2°K in transverse and in longitudinal fields up to 100 kg.

537.311.33 925

Electric Conduction in Semiconductors—H. Fröhlich and G. L. Sewell. (*Proc. Phys. Soc.*, vol. 74, pp. 643–647; November 1, 1959.) In semiconductors with low mobility the usual theory of conduction is invalid. An expression for mobility is derived which should apply in such cases under certain conditions.

537.311.33 926

Volume and Surface Recombination of Injected Carriers in Cylindrical Semiconductor Ingots—J. P. McKelvey. (IRE TRANS. ON ELECTRON DEVICES, vol. ED-5, pp. 260–264; October, 1958. Abstract, PROC. IRE, vol. 47, p. 495; March, 1959.)

537.311.33 927

Metal-to-Semiconductor Contacts: Injection or Extraction for Either Direction of Current Flow—N. J. Harrick. (*Phys. Rev.*, vol. 115, pp. 876–882; August 15, 1959.) Infrared radiation is used to measure the current-density/added-carrier-density characteristics. The nature of the semiconducting surface rather than the metal is the major factor controlling the contact characteristics. Two unusual classes of results are described.

537.311.33 928

Compound Semiconductors—D. A. Wright. (*Electronic Engrg.*, vol. 31, pp. 659–665; November, 1959.) After a general description of some factors influencing performance, applications of individual binary and ternary compounds are considered with reference to tabulated data on their properties. The optimum requirements for thermoelectric refrigeration are noted.

537.311.33 929

On the Study of the Semiconductor Properties of Chalcogenides of Aluminium, Gallium and Indium—E. Kauer and A. Rabenau. (*Z. Naturforsch.*, vol. 13a, pp. 531–536; July, 1958.) Optical measurements confirm that all chalcogenides of Al, Ga and In of the type $A_2^{III}B_3^{VI}$ have semiconductor properties. Band gaps and their temperature dependence are given and compared with the results of other authors.

537.311.33:061.3 930

1958 International Conference on Semiconductors—(*J. Phys. Chem. Solids*, vol. 8, pp. 1–552; January, 1959.) The text is given of papers presented at the conference held in Rochester, N.Y. August 18–22, 1958.

537.311.33:534.2–8 931

The Conductivity of Semiconductors in an Ultrasonic Field—V. A. Zhuravlev and M. I. Kozak. (*Zh. Eksp. Teor. Fiz.*, vol. 36, pp. 343–344; January, 1959.) Measurements have been made on Se, CdS, PbS, Cu_2O, SnO_2 and Ge semiconductors in a 600-kc field of intensity 10 watts/cm^2. Conductivity changes occurred in all cases and are attributed to thermal effects of the irradiation.

537.311.33:538.63 932

The Influence of the Geometry on the Transverse Magnetoresistance Effect in Rectangular Semiconductor Plates—H. J. Lippmann and F. Kuhrt. (*Z. Naturforsch.*, vol. 13a, pp. 462–474; June, 1956.) The resistance effect controlled by the geometry of the specimen is calculated for small and large Hall angles. The theoretical results obtained accord with measurements made on InSb specimens by Weiss and Welker (143 of 1955).

537.311.33:538.632 933

The Influence of the Geometry on the Hall Effect in Rectangular Semiconductor Plates—H. J. Lippman and F. Kuhrt. (*Z. Naturforsch.*, vol. 13a, pp. 474–483; June, 1958.) *See* also 932 above.

537.311.33:539.12.04 934

Temperature-Dependent Defect Production in Bombardment of Semiconductors—G. K. Wertheim. (*Phys. Rev.*, vol. 115, pp. 568–569; August 1, 1959.) The decreased defect density observed after low-temperature bombardment may be due to the production of a metastable vacancy—interstitial pair which may either anneal or form the defect. A temperature dependence in the production rate arises if these two competing processes have different activation energies.

537.311.33:546.26–1 935

Electrical Resistivity Changes Observed in a Semiconducting Diamond after Heat Treatment—R. E. Mutch and F. A. Raal. (*Nature*, vol. 184, suppl. no. 24, pp. 1857–1858; December 12, 1959.) Resistivity measurements on a diamond slab consisting of a blue and a white portion show that both portions are semiconducting to different degrees and are separated by a well-defined boundary. The reduction in donor concentration in both portions on heating appears to be consistent with a bimolecular annealing process, the white portion, with more donor centers, showing a greater change in resistivity.

531.311.33:546.28 936

Infrared Spectrum and Carrier Mobility in the Intrinsic-Conduction Range of Silicon—F. R. Kessler and J. Schnell. (*Z. Naturforsch.*, vol. 13a, pp. 458–461; June, 1958.) Measurements of conductivity and infrared absorption between 1.0 and 3.5μ were made in the temperature range 300°–1000 °K on single-crystal *i*-type Si. The carrier density obtained is in agreement with that derived from Hall-effect measurements by other authors.

537.311.33:546.28 937

Deep Energy Levels in Silicon—H. Irmler. (*Z. Naturforsch.*, vol. 13a, pp. 557–559; July, 1958.) The recombination-center and acceptor energy levels were determined from measurements on pure Si and Si specimens doped with Cu, Ag or Co.

537.311.33:546.28 938

The Control of Resistivity in Pulled Silicon Crystals—A. Trainor and P. T. Harris. (*Proc. Phys. Soc.*, vol. 74, pp. 669–670; November 1, 1959.) This is achieved by pulling with a reduced gas pressure to permit evaporation of the impurity.

537.311.33:546.28 939

Diffusion of Oxygen in Silicon—R. A. Logan and A. J. Peters. (*J. Appl. Phys.*, vol. 30, pp. 1627–1630; November, 1959.) By studying the processs of the donor fomation by heat treatment in crystals of known oxygen concentration, the solubility and diffusivity of oxygen in Si in the temperature range 1250°–1400°C have been determined.

537.311.33:546.28 940

Infrared Studies of Birefringence in Silicon—S. R. Lederhandler. (*J. Appl. Phys.*, vol. 30, pp. 1631–1638; November, 1959.) Techniques are described for observing and measuring permanent and elastic strains.

537.311.33:546.289 941

Zeeman Effect on Light due to Intrinsic Recombination in Germanium—C. Benoit à la Guillaume and O. Parodi. (*J. Electronics Control*, vol. 6, pp. 356–358; April, 1959. In French.) Observations of the Zeeman effect on recombination radiation in As-doped Ge show the energy shift in a field of 40,000 oersteds to be about 4×10^{-4} ev and confirm the hypothesis that this radiation is due to exciton annihilation. See 3369 of 1959 (Gosnet, *et al.*).

537.311.33:546.289 942

Light Emission and Noise Studies of Individual Microplasmas in Silicon *p-n* Junctions—A. G. Chynoweth and K. G. McKay. (*J. Appl. Phys.*, vol. 30, pp. 1811–1813; November, 1959.) "At low currents in the prebreakdown region of broad-area diffused silicon *p-n* junctions in which the breakdown is by an avalanche mechanism, only a few light-emitting microplasmas are present. These appear in succession as the current is increased and the appearance of each spot is accompanied by its own set of characteristic microplasma current pulses. It is found also that effectively all the emitted light arises at these microplasmas and that they carry, essentially, all of the breakdown current. The light intensity of an individual microplasma is roughly proportional to the current flowing through it."

537.311.33:546.289 943

Visible Light from a Germanium Reverse-Biased *p-n* Junction—J. T. Nelson and J. C. Irvin. (*J. Appl. Phys.*, vol. 30, p. 1847; November, 1959.) A phenomenon similar to that observed in Si [see, *e.g.*, 3514 of 1958 (Cheynoweth and Pearson)] is described.

537.311.33:546.289 944

Observation of Direct Tunnelling in Germanium—J. V. Morgan and E. O. Kane. (*Phys. Rev. Lett.*, vol. 3, pp. 466–468; November 15, 1959.) The transition from "indirect" to "direct" tunnelling has been observed experimentally at 77°K. Calculations are compared with experimental data.

537.311.33:546.289:534.2–8 945

Temperature Dependence of Fractional Velocity Changes in a Germanium Single Crystal—F. Stein, N. G. Einspruch, and R. Truell. (*J. Appl. Phys.*, vol. 30, pp. 1756–1758; November, 1959.) Report of interferometer measurements of elastic constants for compressional-wave propagation in the frequency range 30–170 mc.

537.311.33:546.289:535.514.2–15 946

Polarization of Infrared Radiation by Reflection from Germanium Surfaces—D. F. Edwards and M. J. Bruemmer. (*J. Opt. Soc. Amer.*, vol. 49, pp. 860–861; September, 1959.) A Ge surface is an efficient infrared polarizer over a wide range of wavelengths.

537.311.33:546.289:538.63 947

Nernst and Ettingshausen Effect in Germanium between 300 and 750°K—H. Mette, W. W. Gärtner and C. Loscoe. (*Phys. Rev.*, vol. 115, pp. 537–542; August 1, 1959.) Measurements have been made between 300° and 750°K of a) the two effects in single crystals of different conductivity type and with various impurity densities, in magnetic fields of 9000 G; b) the Nernst effect of 2100 G. Results are compared with theory.

537.311.33:546.289:539.23 948

Electrical Conductivity of Amorphous Vapour-Deposited Germanium Films—L. Reimer. (*Z. Naturforsch.*, vol. 13a, pp. 536–541; July, 1958.) The temperature dependance of resistivity was measured in vacuum in the temperature range 20°–400°C. Comparisons are made with the properties of crystalline films.

537.311.33:546.48'19 949

Cyclotron Resonance in $CdAs_2$—M. J. Stevenson. (*Phys. Rev. Lett.*, vol. 3, pp. 464–466; November 15, 1959.)

537.311.33:546.682'86 950

Thermal Conductivity of Indium Antimonide at Low Temperatures—E. V. Mielczarek and H. P. R. Frederikse. (*Phys. Rev.*, vol. 115, pp. 888–891; August 15, 1959.) Umklapp, isotope and boundary scattering contributions to thermal resistivity were calculated theoretically and subtracted from the measured value of thermal resistivity. The number of point impurities subsequently deduced agrees with that given by electrical measurements.

537.311.33:546.682'86 951

The Energy-Dependence of Electron Mass in Indium Antimonide determined from Measurements of the Infrared Faraday Effect—S. D. Smith, T. S. Moss, and K. W. Taylor. (*J. Phys. Chem. Solids*, vol. 11, pp. 131–139; September, 1959.) Values of dE/dK as a function of k_F for the conduction band of InSb, where k_F is the wave vector at the Fermi surface, have been calculated from the Faraday rotation due to free electrons. Results are in agreement with the band structure calculated by Kane (3156 of 1958).

537.311.33:546.682'86 952

Plastic Deformation of InSb by Uniaxial Compression—J. J. Duga, R. K. Willardson, and A. C. Beer. (*J. Appl. Phys.*, vol. 30, pp. 1798–1803; November, 1959.) Plastic deformation by unixaial compression was found to decrease electron mobil ty and magnetoresistance but to have no effect on the Hall coefficient.

537.311.33:546.873'241 953

The Magnetic Susceptibility of Bismuth Telluride—R. Mansfield. (*Proc. Phys. Soc.*, vol. 74, pp. 599–603; November 1, 1959.) The susceptibility has been measured over the temperature range 100°–600°K; the magnitude of the contributions to the susceptibility from various sources is discussed.

537.311.33:548.0 954

The Effect of Pressure on Zinc Blende and Wurtzite Structures—A. L. Edwards, T. E. Slykhouse, and H. G. Drickamer. (*J. Phys. Chem. Solids*, vol. 11, pp. 140–148; September, 1959.)

537.311.33:621.382 955

Temperature Dependence and Lifetime in Semiconductor Junctions—D. A. Jenny and J. J. Wysocki. (*J. Appl. Phys.*, vol. 30, pp. pp. 1692–1698; November, 1959.) The temperature dependence of *p-n*-junction current can be held to a minimum over a given temperature range by using material with a minority-carrier lifetime below a certain value. Calculations applied to GaAs show that an optimum lifetime should be attainable in practice by controlled doping with recombination-center impurities. The upper operating-temperature limit of junction devices is calculated for Ge, Si, InP and GaAs.

537.311.33:621.391.822 956

Load-Dependent Noise in Oxide Semiconductors at Low Frequencies—K. Leberwurst. (*NachrTech.*, vol. 8, pp. 568–580; December, 1958.) The changes of noise characteristics with temperature and oxygen pressure are investigated for a number of sintered oxide semiconductors. From consideration of the mechanism of chemisorption a model is developed for a polycrystalline oxide semiconductor which allows for different types of noise to originate in the interior and at the boundary of unit cells making up the substance. Results of measurements are discussed with reference to an equivalent circuit based on the model.

537.312.8 957

Influence of Collective Effects on the Magnetoresistance of Metals—R. A. Coldwell-Horsfall and D. ter Haar. (*Phys. Rev.*, vol. 115, pp. 891–893; August 15, 1959.) Two methods are used to investigate the influence of the correlations between carriers on the magnetoresistance β. The change in β in both a one-band and a two-band model is considered.

537.533 958

On the Question of Electron Emission of Nickel in the Curie Region—F. Fraunberger and A. Kellerer. (*Z. Phys.*, vol. 154, pp. 419–422; April 7, 1959.) Electron emission and, in particular, exo-electron emission of a Ni specimen was investigated to discover a possible anomaly in the Curie region. No anomaly was found.

537.533:546.431 959

Effect of Oxygen on the Work Function of Barium—P. A. Anderson and A. L. Hunt. (*Phys. Rev.*, vol. 115, pp. 550–552; August 1, 1959.) The effect is measured for various exposures to oxygen.

537.533:546.59 960

Work Function of Gold—P. A. Anderson. (*Phys. Rev.*, vol. 115, pp. 535–554; August 1, 1959.) The measured value of contact difference of potential Au-Ba is found to be 2.31 ±0.02 volts and the work function of Au 4.83±0.02 ev.

537.533:546.621 961

Electron Emission from Plastically Strained Aluminum—W. D. Von Voss and F. R. Brotzen. (*J. Appl. Phys.*, vol. 30, pp. 1639–1645; November, 1959.) An experimental investigation is made of the effect of mechanical variables on emission, both during and after straining. A model is proposed correlating these in terms of the formation of point discontinuities in the surface oxide layer and this is compared with the experimental results.

538.214:546.31 962

Magnetic Susceptibility of Alkali Elements—K. Venkateswarlu and S. Sriraman. (*Z. Naturforsch.*, vol. 13a, pp. 451–458; June, 1958. In English.)

Part 1: Sodium and Potassium (pp. 451–455).

Part 2: Liquid Alloys of Sodium and Potassium (pp. 455–458).

538.22:537.311.31 963

Magnetic Susceptibility and Electrical Resistivity of Au-Mn Alloys—A. Giansoldati, J. O. Linde, and G. Borelius. (*J. Phys. Chem., Solids*, vol. 11, pp. 46–54; September, 1959) Report of measurements on alloys containing 10–75 atomic per cent Mn.

538.221 946

The Magnetic Law of Approach in Plastically Deformed Nickel and Nickel-Cobalt Single Crystals—H. Kronmüller. (*Z. Phys.*, vol. 154, pp. 574–600; April 21, 1959.) Experimental investigation of the influence of plastic deformation at room temperature and at −183° C on the approach to ferromagnetic saturation.

538.221 965

Magnetic Viscosity in Iron due to Carbon Atoms Anchored in Dislocations—G. Biorci, A. Ferro, and G. Montalenti. (*J. Appl. Phys.*, vol. 30, pp. 1732–1735; November, 1959.)

538.221:539.234:537.312.8 966

The Determination of the Curie Temperature of Thin Films by means of the Ferromagnetic Resistance Variation—W. Hellenthal. (*Z. Naturforsch.*, vol. 13a, pp. 566–567; July, 1958.) The dependance of Curie temperature on film thickness is plotted for Ni films. See 1944 of 1959.

538.221:621.318.134 967

The Influence of the Valency States of Cations on Electrical Conductivity of Mg-Mr Ferrite—S. Krupička and Závěta. (*J. Electronics Control*, vol. 6, pp. 333–336; April, 1959.) The resistivity of iron-deficient Mg ferrite is low, owing to the presence of Fe^{2+} ions, but can be greatly increased by the addition of Mn. This is attributed to the formation of stable Mn^{4+}-Fe^{2+} pairs, preventing the Fe^{2+} ions taking part in the conduction process.

538.221:621.318.134 968

Evidence for Triangular Moment Arrangements in M0. Mn$_2$0$_3$—I. S. Jacobs. (*J. Phys. Chem. Solids*, vol. 11, pp. 1–11; September, 1959.) A consequence of the triangular arrangement of moments suggested by Yafet and Kittel (3414 of 1952) is that at field strengths higher than those required for ferrimagnetic domain alignment, there should be a linear increase in net magnetization with field. Measurements on several spinel-type compounds show this differential susceptibility.

538.221:621.318.134 969

The Magnetocrystalline Anisotropy of Cobalt-Substituted Manganese Ferrite—R. F. Pearson. (*Proc. Phys. Soc.*, vol. 74, pp. 505–512; November 1, 1959.) The anisotropy due to Co ions in Mr ferrite varies linearly with Co concentration up to 25 per cent; the variation is not as great as in magnetite or as would be expected from Co ferrite values.

538.221:621.318.134 970

Paramagnetic Resonance of Yb^{3+} in Yttrium Gallium Garnet—D. Boakes, G. Garton, D. Ryan, and W. P. Wolf. (*Proc. Phys. Soc.*, vol. 74, pp. 663–665; November 1, 1959.)

538.221:621.318.134 971

Magnetic Properties of Rare Earth Ions in Garnets—W. P. Wolf. (*Proc. Phys. Soc.*, vol. 74, pp. 665–667; November 1, 1959.) *See* also 970 above.

538.221:621.318.134:538.566 972

Faraday Effect in Various Small Ferrite Rods—F. Picherit. (*Compt. rend. Acad. Sci., Paris*, vol. 249, pp. 69–70; July 6, 1959.) The polarimetric analyzer described in 3574 of 1958 has been applied to the study of the Faraday effect in 4B ferrite rods of diameter down to 1.6 mm.

538.221:621.318.134:538.569.4 973

Magnetic Resonance of Ferrites at the Compensation Temperature—J. Paulevé. (*Ann. Télécommun.*, vol. 13, pp. 311–324; November/December, 1958. and vol. 14, pp. 2–20; January/February, 1959.) A study of ferrimagnetic resonance theory at temperatures below the Curie point supported by measurements on Li chrome ferrites and Gd and Er garnets. 43 references.

538.221:621.318.134:538.569.4 974

The Influence of Conducting Surfaces on the Gyromagnetic Resonance of Ferrite Bodies—W. Haken. (*Arch. elekt. Übertragung*, vol. 12, pp. 562–566; December, 1958.) The calculation of demagnetizing factors by magnetostatic methods must be modified if conducting surfaces are present. Using a mirror-image principle the demagnetizing factors are calculated for magnetized ferrite strips in rectangular waveguides; measurement results confirm the analysis.

538.222:538.569.4 975

Optical Detection of Paramagnetic Resonance Saturation in Ruby—I. Wieder. (*Phys. Rev. Lett.*, vol. 3, pp. 468–470; November 15, 1959.) Description of experimental measurements.

538.632:546.87:539.23 976

Hall Effect of Bismuth Films Condensed by Quenching—W. Buckel. (*Z. Phys.*, vol. 154, pp. 474–485; April 7, 1959.)

538.652 977

Measurement of the Linear Magnetostriction of Hard-Worked Nickel—H. E. Stauss. (*J. Appl. Phys.*, vol. 30, pp. 1648–1650; November, 1959.)

539.2:539.12.04 978

The Effects of Nuclear Radiation on Materials—T. P. Flanagan. (*J. Electronics Control*, vol. 6, pp. 337–346; April, 1959.) The effects of gamma and neutron radiation are described. Magnetic materials and thermocouples are included.

621.315.612.015.5 979

Dielectric Breakdown of Porous Ceramics—R. Gerson and T. C. Marshall. (*J. Appl. Phys.*, vol. 30, pp. 1650–1653; November, 1959.) The measured dielectric strength in porous materials is a function of porosity, void size, and of the dimensions of the test specimen.

621.315.615.015.5 980

Conduction Currents in Liquid *n*-Hexane under Microsecond Pulse Conditions—A. H. Sharbaugh and P. K. Watson. (*Nature*, vol. 184, suppl. no. 26, pp. 2006–2007; December 26, 1959.) No evidence for electron multiplication was obtained at fields below 1.25 mv/cm. At fields in the vicinity of breakdown, large values of current density were measured which may lead to local vaporization and the formation of bubbles with subsequent breakdown of the liquid dielectric.

621.315.616.9:621.3.048 981

An Alternative Method for Comparing the Electrical Properties of Epoxy Casting Resins—D. A. Thompson. (*Electronic Engrg.*, vol. 31, pp. 686–687; November, 1959.)

669.21/.22:621.382.3.032.27 982

Investigations on some Noble-Metal Alloys used in Semiconductor Techniques—K. Müller and W. Merl. (*Elektrotech. Z., Edn A*, vol. 80, pp. 515–518; August 1, 1959.) Physical and mechanical properties are tabulated of various alloys of Ag and Au with group-III and group-V elements suitable as contact materials for semiconductor devices, and curves of specific resistance as a function of alloy composition are given.

MEASUREMENTS AND TEST GEAR

529.786+621.3.018.41 (083.7) 983

The Determination of Time and Frequency—(*Tech. Mitt. PTT*, vol. 37, pp. 1–32; January 1, 1959.) The text is given of the following papers presented at a meeting of the Swiss national committee of URSI at Neuchâtel, 1958.

a) **The Scientific Importance of Atomic Frequency Standards**—J. Rossel (pp. 2–6. In French).

b) **The Atomic Clocks**—J. Bonanomi (pp. 6–9).

c) **The Technique of the Atomic Determination of Frequency and Time**—J. De Prins and P. Kartaschoff(pp. 10–14).

d) **Problems of Astronomic Time**—J. P. Blaser, and W. Schuler (pp. 14–17. In French).

e) **Time Signals and Standard Frequencies**—C. Wyser (pp. 18–23. In French).

f) **The Time Service of the Swiss Post, Telegraph and Telephone Administration**—K. J. Bohren (pp. 23–32).

529.786+621.3.018.41(083.74) 984

An Improved Caesium Frequency and Time Standard—L. Essen and J. V. L. Parry.(*Nature*, vol. 184, suppl. no. 23, p. 1791; December 5, 1959.) Preliminary measurements show that a precision of a few parts in 10 inches is practicable. See 204 of 1958.

621.317.331.029.6 985

Measurement of Microwave Resistivity by Eddy-Current Loss in Spheres—T. Kohane and M. H. Sirvetz. (*Rev. Sci. Instr.*, vol. 30, pp. 1059–1060; November, 1959.) Applies to materials in which "skin effect" is appreciable in samples of ordinary size.

621.317.36:621.374.32 986

Digital Rate Synthesis for Frequency Measurement and Control—T. J. Rey. (PROC. IRE, vol. 47, pp. 2106–2112; December, 1959.) The signal whose period is to be measured is in the form of a pulse train of constant repetition rate. This train is compared with the average repetition rate of a series of nonuniformly spaced pulses from a standard source, the differences being used for frequency indication, or in a feedback loop for control purposes.

621.317.373:621.374.4 987

A Phase Multiplier—E. Augustin. (*Hochfrequenz.*, vol. 67, pp. 84–87; November, 1958.) Frequency multiplication and mixing are used in the equipment described to obtain a ten- or hundred-fold increase of very small phase angles at 10 kc. Phase multiplication by a factor of 1000 may be achieved using a carefully designed circuit and highly stable supplies.

621.317.412 988

Application of a Weakly Inhomogeneous Magnetic Field to the Quantitative Determination of Small Ferromagnetic Admixtures in Susceptibility Measurements— D. Gerstenberg. (*Z. Metallk.*, vol. 50, pp. 472–477; August, 1959.) Ferromagnetic content of 10^{-8} gram in a specimen can be determined with the apparatus described.

621.317.412 989

Absolute Measurement of Magnetic Susceptibility by a Quasistatic Induction Method—M. C. Mesnage, P. Grivet and M. Sauzade. (*Compt. rend. Acad. Sci., Paris*, vol. 249, pp. 59–60; July 6, 1959.) Changes in induction when a hollow cylinder placed in the gap of a permanent magnet is filled with a liquid or a gas are observed by a null method. Results of observations on 0_2 and N0 are given.

621.317.742 990

VSWR Indicators with Automatic Read-Out—M. Kollanyi and R. M. Verran. (*Electronic Engrg.*, vol. 31, pp. 666–671; November, 1959.) A probe carriage is moved along a slotted line to produce stored values of the maximum and minimum detected signals. Different methods of obtaining a direct reading of the amplitude ratio of these signals are described with circuit details.

621.317.75:534.44 991

Method of Obtaining Amplitude and Phase Spectra of a Transient Function using Graphical Input Data—W. J. Remillard. (*J. Acoust. Soc. Amer.*, vol. 31, pp. 531–534; April, 1959.) A method of harmonic analysis is describe using commercial equipment in conjunction with switching, adding and triangular-wave shaping circuits.

621.317.77:621.373 992

A System for Providing a Precise Vector Voltage—D. J. Collins and J. E. Smith. (*Electronic Engrg.*, vol. 31, pp. 684–685; November, 1959.) Schematic details are given of a means of producing, from a reference oscillator source, signals of any desired amplitude and phase.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

551.510.62:621.372.413 993

A Microwave Refractometer—R. Schünemann and W. Steffen. (*Hochfrequenz.*, vol. 67, pp. 78–83; November, 1958.) A cavity-resonator-type refractometer operating at 9.4 kmc is described and some initial measurements near the ground are discussed.

621–52:621.318.57:621.387 994

Cold-Cathode Tube Circuits—H. Liebendörfer. (*Electronic Radio Eng.*, vol. 36, pp. 436–442; December, 1959.) Describes briefly various basic control circuits using cold-cathode triodes and their application in systems for lighting control, flame detection in oil-fired boilers, thread-break detectors, etc.

621–52:621.318.57:621.387 995

ACCESS—a Static Switching System using Cold-Cathode Tubes—R. W. Brierley. (*Electronic Engrg.*, vol. 31, pp. 646–654; November, 1959.) A description of some features of an automatic production control system developed by the Austin Motor Co.

621–57:537.228.4 996

Flame-Sprayed Ceramic Dielectric for Transducers—G. V. Planner and A. Foster. (*J. Brit. IRE*, vol. 19, pp. 699–702; November, 1959.) "A high-speed electromechanical transducer system based on the Johnson-Rahbek effect is described, in which a flame-sprayed ceramic of high permittivity replaces the conventional semiconductor."

621.363 997

Thermomagnetic Generator—J. F. Elliott. (*J. Appl. Phys.*, vol. 30, pp. 1774–1777; November, 1959.) Calculations are given for a practical thermal-energy converter using gadolinium. Energy is extracted from heat sources near 20°C; reasonable efficiency and power output are claimed. See 1309 of 1949 (Brillouin and Iskenderian).

621.383.4:535–15:621.397.331.2 998

Electronic Scanning System for Infrared Imaging—M. E. Lasser, P. H. Cholet, and R. B. Emmons. (PROC. IRE, vol. 47, pp. 2069–2075; December, 1959.) An infrared image is focused on a semiconducting window of silicon at one end of a tapered tube. This is scanned by an electron beam creating free carriers locally in the semiconductor and producing a moving opaque spot which absorbs energy as the image passes through the tube to an infrared detector. The detector output is fed to a television-type monitor for display.

621.384.622.21 999

Longitudinal Movement of Electrons and Tolerances in a Linear Accelerator—A. D. Vlasov. (*Radiotekh. Elektron.*, vol. 4, pp. 295–302; February, 1959.) Determination of the injection conditions governing the acceleration and the compactness of the energy spectrum of accelerated particles.

621.385.833 1000

The Construction of Four-Pole Magnetic Lenses without Iron—A. Septier and G. Chartier. (*Compt. rend. Acad. Sci., Paris*, vol. 249, pp. 64–66; July 6, 1959.) A theoretical investigation of the magnetic field due to four-pole linear and sheet currents has been applied to the design of a strong-focusing magnetic lens which has been constructed and found to give satisfactory results.

621.387.4:621.374.3 1001

Fast Neutron Time-of-Flight Spectrometer—G. C. Neilson, W. K. Dawson, and F. A. Johnson. (*Rev. Sci. Instr.*, vol. 30, pp. 963–975; November, 1959.) A detailed description is given of a neutron spectrometer incorporating a simple time/pulse-height converter suitable for use in the mμs region. Either RF beam deflection or the gamma coincidence method is used to provide zero time indication.

621.56:537.322.1 1002

Thermoelectric Cooling—(*Electronic Engrg.*, vol. 31, pp. 690–692; November, 1959.) Refrigeration figures-of-merit of thermocouple elements are shown to be greater for certain semiconductors than for metals. Practical applications of thermoelectric cooling are discussed.

PROPAGATION OF WAVES

621.391.812.62 1003

Correlation of Meteorological Phenomena with Propagation beyond the Horizon over the Mediterranean—L. Bonavoglia. (*Alta Frequenza*, vol. 27, pp. 815–824; December, 1958.) Results of propagation tests between Sardinia and Minorca (see also 2229 of 1958) and of meteorological measurements are compared.

621.391.812.62 1004

Path Antenna Gain in an Exponential Atmosphere—W. J. Hartman and R. E. Wilkerson. (*J. Res. Nat. Bur. Stand.*, vol. 63D, pp. 273–286; November/December, 1959.) This extension of previous estimates of the loss in gain caused by tropospheric or stratospheric scattering includes the effect of an exponential model atmosphere and a scattering cross section inversely proportional to the fifth power of the scattering angle. Assuming isotropic turbulence the loss is about 2-db less than Staras' estimate (235 of 1959), which ignored the exponential atmosphere. The loss would be further reduced if anisotropic turbulence were included.

621.391.812.621 1005

Effect of Atmospheric Horizontal Inhomogeneity upon Ray Tracing—B. R. Bean and B. A. Cahoon. (*J. Res. Nat. Bur. Stand.*, vol. 63D, pp. 287–292; November/December, 1959.) The effect is negligible above 1 km. Below this, rays at elevation angles less than about 2° are sensitive to extreme horizontal inhomogeneities, as in ducting conditions. The method is partly graphical and unjustifiable for routine correction of ray paths.

621.391.812.621 1006

Partial Reflections in the Atmosphere and Long-Distance Propagation: Parts 1 and 2—F. du Castel, P. Misme, and J. Voge. (*Ann. Télécommun.*, vol. 13, pp. 209–214 and pp. 265–270; July/August and September/October, 1958.) Refractive-index measurements show the existence in the atmosphere of small layers (feuillets) each with a refractive index different from that of the mean surrounding value. Interpretation of the results suggests that the atmosphere may be considered as a dielectric medium composed of such layers which may be stable or unstable and identifiable by the slope of the index gradient through the layer.

621.391.812.63 1007

A Discussion of Ionospheric Demodulation Near Gyro-Frequency—G. L. Goodwin. (*Aust. J. Phys.*, vol. 12, pp. 157–163; June, 1959.) The effect occurs at a height of about 90 km and does not appear to decrease at dawn. A wave reflected from the F-layer is demodulated by unequal amounts during its two passages through the region. The results cannot be explained by the theory of wave interaction.

621.391.812.8 1008

Comparison of the Long-Term Prediction Methods of C.R.P.L. and S.P.I.M.—P. Halley, D. Lepechinsky and P. Mouchez. (*Ann. Télécommun.*, vol. 13, pp. 254–264; September /October, 1958.)

RECEPTION

621.391.812.3:621.317.6.087.4 1009

Correlation between Fading Signals—J. Bell. (*Electronic Tech.*, vol. 37, pp. 36–40; January, 1960.) An instrument is described for the determination of the correlation coefficient between two positive rectified fading signals using a sum-and-difference method.

621.391.821:621.3.087.4 1010

Recording of the Mean Square Value of Atmosphereic Noise—F. Carbenay. (*Compt. rend. Acad. Sci., Paris*, vol. 249, pp. 67–68; July 6, 1959.) The method used at the CNET is described. A vertical omnidirectional antenna above a metal earth is used and the square of the intensity of the linearly amplified current at a given frequency is measured by three thermocouples of different sensitivity each associated with a galvanometer. Recordings may be made photographically.

STATIONS AND COMMUNICATION SYSTEMS

621.391 1011

Poisson, Shannon, and the Radio Amateur—J. P. Costas. (PROC. IRE, vol. 47, pp. 2058–2068; December, 1959.) From a statistical analysis of the problem of communicating in a congested band of frequencies, it is concluded that reducing the required bandwidth for a service, in the hope of acquiring an exclusive channel does not always offer the best prospect of intelligible reception. The use of wider bandwidths is advocated, particularly for amateur work where specific frequencies are not allocated for individual transmitters, and for military situations in which considerable interference, unintentional or deliberate, may be expected.

621.391 1012

Error-Correcting Codes—a Linear Programming Approach—E. J. McCluskey, Jr. (*Bell Syst. Tech. J.*, vol. 38, pp. 1485–1512; November, 1959.) Theorems are given for the matrices used to construct systematic error-correcting codes and are proved for minimum-redundancy codes.

621.391:534.75 1013

Indices of Signal Detectability Obtained with Various Psychophysical Procedures—J. A. Swets. (*J. Acoust. Soc. Amer.*, vol. 31, pp. 511–513; April, 1959.) See also 3854 of 1959 (Tanner and Birdsall).

621.391:534.78 1014

Communication Efficiency of Vocoders—A. R. Billings. (*Electronic Radio Eng.*, vol. 36, pp. 449–453; December, 1959.) A vocoder, in which speech is modulated and a single sideband passed to the analyzer filters through an amplitude limiter, is shown on comparison with the convention vocoder to have a greater communication efficiency, this being defined as the ratio of the actual rate of information to the rate of transmission in an ideal system subject to the same restrictions.

621.391:621.376.5 1015

Ideal Binary Pulse Transmission by A.M. and F.M.—E. D. Sunde. (*Bell Syst. Tech. J.*, vol. 38, pp. 1357–1426; November, 1959.) Intersymbol interference can be avoided in binary pulse transmission by FM without the need for a wider channel bandwidth than in DSB AM for equal pulse transmission rates. Explicit general expressions are derived for the appropriate shaping of the band-pass channel. The optimum division of channel shaping between transmitting and receiving filters for FM and AM with random noise, and the relation between error probability and signal/noise ratio are discussed.

621.391.822 1016

Transition Probability Densities of the Smoothed Random Telegraph Signal—W. M. Wonham. (*J. Electronics Control*, vol. 6, pp. 376–384; April, 1959.) An extension of earlier analysis (3634 of 1958) to derive the transition probability functions of the associated Markov process.

621.395:621.372 1017

The South Lancashire Radiophone Service—(*Brit. Commun. Electronics*, vol. 6, pp. 858–861; December, 1959.)

Part 1—The General System—L. T. Arman (pp. 858–860). A VHF mobile radio service provides communication between land mobile stations and the public telephone network.

Part 2—The Radio-Telephone Equipment—B. Armstrong (pp. 860–861). A discussion of the technical specification.

621.395:621.376.56 1018

A Delta Modulation System using Junction Transistors—B. E. Williams. (*Electronic Engrg.* vol. 31, pp. 674–680; November, 1959.) A speech link operating at a digit frequency of 14 kc uses alternating binary encoding—a form in which the signal has no dc component—and requires less bandwidth than normal binary code.

621.395.665.1:534.861 1019

Contribution on the Reciprocal Control of Dynamic Range—H. Völz. (*Hochfrequenz.*, vol. 67, pp. 87–94; November, 1958.) Six-terminal networks are introduced in this theoretical consideration of dynamic compression and restitution of the original dynamics after the signal has passed through a noisy communication channel. Practical systems designed to avoid the occurrence of transients are discussed.

621.396.65 1020

Convention on Radio Links—(*Alta Frequenza*, vol. 27, pp. 578–832; December, 1958.) Third issue covering the proceedings of a convention held in Rome, June 5–8, 1957. Second issue: 2014 of 1959. Abstracts of some of the papers are given individually; titles of others are as follows:

a) **Microwave Amplification by Earthed-Grid Triodes**—G. B. Stracca (pp. 578–614).
b) **Applications of Ferrites in Microwave Radio Links**—M. Vadnjal (pp. 615–628).
c) **Frequency Control of Klystrons by Reference Cavity**—L. Del Bello (pp. 629–633).
d) **On the Improvement of Linearity of Reflex Klystrons used as Frequency-Modulated Oscillators**—G. Cicconi (pp. 634–682.)
e) **New Components for Miniaturized Microwave Radio Links**—M. Mueller (Müller) (pp. 683–698. In English).
f) **Problems of Automatic Switching**—A. Ricagni (pp. 699–715).
g) **Criteria and Measurements of Suitability of Equipment for Radio Links**—G. Monti-Guarnieri (pp. 716–751).
h) **A 6000-Mc/s Communication System for 300 Telephone Channels**—E. Viti (pp. 752–778).
i) **Multiplex for Three High-Quality Music Channels**—A. Pasini (pp. 779–789).
j) **Directive Aerials with Dual Polarization**—G. Parmeggiani (pp. 794–814).

621.396.74.029.51/.53(43) 1021

Long and Medium-Wave Transmitters in the German Federal Republic—(*Rundfunktech. Mitt.*, vol. 2, pp. 304–307; December, 1958.) *See* also 1023 below.

621.396.74.029.55(43) 1022

Short-Wave Frequencies in the German Federal Republic—(*Rundfunktech. Mitt.*, vol. 2, p. 303; December, 1958.)

621.396.74.029.62(43) 1023

List of VHF Transmitters in the German Federal Republic—(*Rundfunktech. Mitt.*, vol. 2, pp. 300–303; December, 1958.) The lists and map of transmitter locations give the position as of November 1, 1958.

621.396.931 1024

Radio Telephony in French Express Trains—[*Engineer* (*London*), vol. 208, pp. 158–159; August 28, 1959.)] On the route Paris-Lille a VHF link operating at 160 mc is provided between trains and a chain of 11 lineside stations with a cable connection to a Paris exchange.

SUBSIDIARY APPARATUS

621.3.087.4:621.396.96 1025

Electronic Line-Storage Device for the Compression of the Frequency Band of Periodically Recurring Signals, particularly of Radar Displays—K. Lange. (*Nachrichtentech. Z.*, vol. 11, pp. 619–627; December, 1958.) The operation of a CR -tube line-storage system is discussed. Equipment is described for measuring the integrating characteristics of such storage devices. See also 3744 of 1956 (Meinke and Groll).

621.316.722.027.5:621.385.4 1026

A Small, High-Voltage, Regulated Power Supply with Variable Output—J. D. O'Toole. (*Electronic Engr.*, vol. 31, pp. 681–683; November, 1959.) The design of a regulating circuit operating in the range 1–4kv is described which uses a beam tetrode as a series regulator and two double triodes in the feedback circuit.

621.316.722.073.3:621.385.032.213 1027

A Transistorized L. T. Regulator—J. H. Deichen. (*Electronic Engr.*, vol. 31, pp. 688–689; November, 1959.) "The methods of providing a stabilized dc supply for tube heaters are examined and their limitations are discussed. A simple circuit using a transistor and two Zener diodes is then given and its design and performance described."

TELEVISION AND PHOTOTELEGRAPHY

621.397:621.391.8.029.63 1028

Some Aspects of Television Reception on Band V—H. N. Gant. (*J. Brit. IRE*, vol. 19, pp. 727–733; November, 1959.) Band-V television receiver design problems are considered with regard to available components. Suitable antenna systems and their performance figures are noted.

621.397.132.001.4:621.317.755 1029

A Vectorscope Unit—K. G. Freeman. (*Electronic Engrg.*, vol. 31, pp. 655–658; November, 1959.) Circuit details are given of an instrument which uses the principles of coherent detection to provide a vectorial display of NTSC subcarrier chrominance signals.

621.397.232.2:621.391.82 1030

Co-channel Television Interference and its Reduction—E. W. Chapin, L. C. Middlekamp, and W. K. Roberts. (IRE TRANS. ON BROADCAST TRANSMISSION SYSTEMS, vol. PGBTS-10, pp. 3–24; June, 1958. Abstract, PROC. IRE, vol. 46, p. 1665; September, 1958.)

621.397.232.2:621.391.82 1031

Reduction of Co-channel Television Interference by Very Precise Offset Carrier Frequency—L. C. Middlekamp. (IRE TRANS. ON BROADCAST TRANSMISSION SYSTEMS, vol. PGBTS-12, pp. 5–10; December, 1958.) The effect on picture degradation of frequency-offset variation is investigated, and the improvement in picture quality to be gained by using precise offset values held to a close tolerance is discussed.

621.397.232.2:621.391.82 1032

Investigation of the Operation of Television Transmitters with Precise Offset-Carrier Frequencies—H. Hopf. (*Rundfunktech. Mitt.*, vol. 2, pp. 265–276; December, 1958.) The reduction of the transmitter-frequency tolerance to about ±2.5 cps results in an improvement by more than 10 db in the RF protection ratio (*see* also 1030 above). Curves of protection ratios are plotted for various types of carrier-offset and include a set of curves of recommended ratios for the 625-line CCIR system, with a reference-interference ratio of 30 db for carrier offset equal to $\frac{2}{3}$× line frequency.

621.397.334 1033

A New Approach to Kinescope Beam Convergence—J. W. Schwartz and P. E. Kaus. (IRE TRANS ON ELECTRON. DEVICES, vol. ED-5, pp. 275–282; October, 1958. Abstract, PROC. IRE, vol. 47, p. 495; March, 1959.)

621.397.334:621.374.33 1034

A Gating Circuit for Single-Gun Colour Television Tubes.—K. G. Freeman. (*J. Brit. IRE*, vol. 19, pp. 667–677; November, 1959.) Gating circuits and their limitations are discussed. A new gating circuit is described, which employs low-level gating of the red, green and blue video signals in conjunction with a wide-band amplifier.

621.397.6.001.4 1035

Laboratory Television Transmitter for the Intermediate-Frequency Range—P. Klopf. (*Rundfunktech. Mitt.*, vol. 2, pp. 253–264; December, 1958.) A modulator unit is described for use in tests on residual-sideband transmission for the 625-line CCIR system. The quality of transmission via the modulator and a standard IF receiver is examined.

621.397.61:535.88 1036

Further Development of the Television Episcope—R. Theile and F. Pilz. (*Rundfunktech. Mitt.*, vol. 2, pp. 290–292; December, 1958.) Details are given of an improved version of the equipment described in 1374 of 1959.

621.397.7—182.3 1037

A Contribution to the Design and Construction of Television Outside-Broadcast and Film-Recording Vehicles—G. Schadwinkel and H. Käding. (*Rundfunktech. Mitt.*, vol. 2, pp. 277–289; December, 1958.) Vehicles developed by the West German radio organizations and their equipment are described.

621.397.72 1038

Some Aspects of the Design of a Small Television Station—A. Harris. (*J. Brit. IRE*, vol. 19, pp. 705–721; November, 1959.) "Some of the factors governing the choice of equipment for use in small commercial television stations in isolated areas are discussed. Specific reference is made to two different types of installation at station ZBM-TV, Bermuda, using the same basic equipment."

621.397.74(43) 1039

The Television Network of the German Federal Republic—(*Rundfunktech. Mitt.*, vol. 2, pp. 293–299; December, 1958.) Tabulated data on television transmitters as of November 1, 1958, with maps showing their location and that of television links.

621.397.743 1040

Climatic Resistance, Operational Reliability, and Emergency Circuitry of Television Frequency Translators—K. Fischer, H. J. Fraisse, and W. Marks. (*Telefunken Ztg.* vol. 32, pp. 47–58, March, 1959. English summary, pp. 73–74.) The effects of outdoor operation on unattended television relay equipment are discussed. Means of minimizing these effects and of ensuring the maintenance of services are illustrated by examples of specially designed units.

TRANSMISSION

621.396.61:621.372.51 1041

Gauging Transmitters to Overcome Jamming—P. W. Esten. (*Electronics*, vol. 32, pp. 68–70; November 27, 1959.) A combining network enables four transmitters to feed a common antenna to increase radiated power.

621.396.61:621.391.82 1042

Interference Area of a Transmitter for Tropospheric-Scatter Radio Links—A. Chinni (*Alta Frequenza*, vol. 27, pp. 825–832; December, 1958.) A nomogram has been constructed for obtaining the distance from a transmitter at which the interference field intensity exceeds a certain level for 0.1 per cent of the time, for any given frequency, radiated power, antenna gain and radiation pattern.

TUBES AND THERMIONICS

621.382.2 1043

Zener Diodes—their Properties and Applications—J. M. Waddell and D. R. Coleman. (*Wireless World*, vol. 66, pp. 17–21; January, 1960.)

621.382.2:621.391.822 1044

The Noise of Diodes—W. Drechsel. (*NachrTech.*, vol. 8, pp. 538–541; December, 1958.) Fundamental causes of LF and HF noise in semiconductor diodes are considered and the choice of diode characteristics for optimum performance in static, dynamic, and mixer operation is discussed.

621.382.2/.3:539.12.04 1045

How Radiation Affects Semiconductor Devices—M. F. Wolff. (*Electronics*, vol. 32, pp. 55–57; November 27, 1959.) Transistor and diode damage and the change in circuit characteristics due to radiation effects are noted. Radiation-resistant devices such as the thin-base Ge transistor and the tunnel diode are mentioned.

621.382.23 1046

The Tunnel Diode—Circuits and Applications—I. A. Lesk, N. Holonyak, Jr. and U. S. Davidsohn. (*Electronics*, vol. 32, pp. 60–64; November 27, 1959.) Details are given of the characteristics of tunnel diodes possessing a high negative conductance and a high resistance to nuclear-radiation and temperature effects. Their applications in amplifier, oscillator and detector circuits are illustrated.

621.382.23:621.372.44 1047

Surface-Dependent Losses in Variable-Reactance Diodes—D. E. Sawyer. (*J. Appl. Phys.*, vol. 30, pp. 1689–1691; November, 1959.) High-frequency performance is lowered by these losses which cause the series equivalent resistance to be larger than the integrated bulk resistance. The effect is shown to vary with surrounding atmosphere and an explanatory model is given.

621.382.3:621.391.822 1048

The Noise Equivalent Circuit of the Transistor—G. Winkler. (*NachrTech.*, vol. 8, pp. 542–548; December, 1958.) The representation of transistor noise in the form of an equivalent noise three-pole network using noise parameter measurements or by an equivalent circuit based on calculation is described.

621.382.3.012.6 1049

Transistor Matching Impedances—A. G. Bogle. (*Electronic Tech.*, vol. 37, pp. 28–30; January, 1960.) Calculations are described relating the variation of input impedance, output impedance and gain of a Type-OC 71 transistor to changes of operating point and negative feedback.

621.382.3.032.27 1050

Application of a Photogravure Process to the Deposition of Metal Contacts on Semiconductors—P. Michelet and J. Vareine. (*Onde élect.*, vol. 39, pp. 858–862; November, 1959.) A photosensitive synthetic resin is used in the process which is applied to HF Si transistors. Reference is made to the elimination of the superficial oxide layer by the process.

621.382.333 1051

An 85-Watt-Dissipation Silicon Power Transistor—R. W. Aldrich, R. H. Lanzl, D. E. Maxwell, J. O. Percival, and M. Waldner. (IRE Trans. on Electron Devices, vol. ED-5, pp. 211–215; October, 1958. Abstract, Proc. IRE, vol. 47, p. 494; March, 1959.)

621.382.333 1052

Large-Signal Rise Times in Junction Transistors—W. W. Gärtner. (IRE Trans. on Electron Devices, vol. ED-5, p. 316; October, 1958.) An investigation of collector current response using a more exact expression for α_N than that used by Moll (885 of 1955) yields considerably higher values for the large-signal rise times.

621.382.333 1053

The Physical Interpretation of Measurements on Transistors—F. J. Hyde (*J. Electronics Control*, vol. 6, pp. 362–364; April, 1959.) The equation for the effective lifetime of holes in the base given by Deb and Daw (2043 of 1959) is shown to be not generally valid, and a modified equation is derived.

621.382.333:621.317.3 1054

Measurement of the High-Frequency Base Resistance and Collector Capacitance of Drift Transistors—F. J. Hyde and T. E. Price. (*J. Electronics Control*, vol. 6, pp. 347–355; April, 1959.) A modified Turner circuit (354 of 1955) is used to determine $r_{bb'}$ and c_c separately. The effect of stray emitter-collector capacitance on the measurements is analyzed.

621.382.333:621.318.57 1055

Two-Terminal Asymmetrical and Symmetrical Silicon Negative-Resistance Switches—R. W. Aldrich and N. Holonyak, Jr. (*J. Appl. Phys.*, vol. 30, pp. 1819–1824; November, 1959.) "By making use of an emitter region shorted by a metallic contact to an adjacent base region, a new form of *p-n-p-n* switch is obtained. Several new structures are described, including a symmetrical (or ac) switch. Typical experimental results on switches which break down in the range from 25 to 40 volts are presented."

621.382.333.33 1056

A Silicon *p-n-p-n* Power Triode—K. Kawana and T. Misawa. (*J. Electronics Control*, vol. 6, pp. 324–332; April, 1959.) A simplified theory assuming equal ionization for electrons and holes is presented which explains the controlling effect of base current on the *I/V* charactieristics of the triode. Characteristics of a developmental triode are given.

621.382.333.4 1057

The Emitter Tetrode—R. A. Gudmundsen. (IRE Trans. on Electron Devices, vol. ED-5, pp. 223–225; October, 1958. Abstract, Proc. IRE, vol. 47, p. 495; March, 1959.)

621.383.4 1058

The Limiting Sensitivity and the Noise Spectrum of Germanium Junction Photodiodes—L. Ya. Pervova. (*Radiotekh. Elektron.*, vol. 4, pp. 330–334; February, 1959.) The sensitivity threshold was of the order of 10^{-10} lumen second $^{1/2}$ with a light-beam modulation frequency 70 cps. The noise spectrum in the range 0.5 cps to several hundres of cps obeys the $1/f$ law; for higher frequencies the noise is close to the shot noise of the photodiode dark current.

621.385 (083.72) 1059

Vacuum Diodes and Grid-Controlled Vacuum Tubes: Terminology—(*Nachrichtentech. Z.*, vol. 11, pp. 635–644; December, 1958.) A list of 138 terms and their definitions recommended by the Nachrichtentechnische Gesselschaft; the corresponding English and French terms are given in most cases. This is an amended version of 652 of 1958.

621.385.1:621.391.822 1060

Flicker Noise of Valves in the L.F. Region—H. Mutschke. (*NachrTech.*, vol. 8, pp. 585–590; December, 1958.) The mechanism of flicker noise is discussed with reference to previous work [*e.g.*, 1284 of 1957 (Lindemann and van der Ziel)] and results are given of tests on various types of tubes to determine the dependence of flicker noise on heater voltage, anode current and running time.

621.385.3.029.63 1061

PC86, a Newly Developed Grounded-Grid Triode for Decimetre Waves (U.H.F.)—H. Bünger. (*Telefunken Ztg.*, vol. 31, pp.262–265; December, 1958. English summary, p. 274.) The glass tube described is designed for operation at frequencies up to 800 mc; its amplification factor is 68, and its slope 14 ma/volt.

621.385.6 1062

Periodic Electrostatic Focusing of a Hollow Electron Beam—C. C. Johnson. (IRE Trans. on Electron Devices, vol. ED-5, pp. 233–243; Abstract, Proc. IRE, vol. 47, p. 495; March, 1959.)

621.385.6 1063

A Symmetry Property of Space-Charge Waves in Accelerated Electron Beams—I. P. Shkarofsky. (IRE Trans. on Electron Devices, vol. ED-5, pp. 283–288; October, 1958. Abstract, Proc. IRE, vol. 47, p. 495; March, 1959.)

621.385.6 1064

On Speed Charge Waves—R. H. C. Newton. (*J. Electronics Control*, vol. 6, pp. 321–323; April, 1959.) Comment on 2058 of 1959 (Trevena).

621.385.6 1065

Calculation of Electron-Beam Divergence at Medium Gas Pressures—B. H. Wadia. (*J. Electronics Control*, vol. 6, pp. 307–320; April, 1959.) A method is described involving step-by-step integration, by which the beam profiles in the presence of gas at $10^{-6}-10^{-7}$ mm Hg can be estimated. Universal curves are given which enable the designer to estimate the modification of the "perfect-vacuum" behavior ($<10^{-8}$ mm Hg) to be expected in practical tubes.

621.385.6:621.375.9 1066

The Kinetic Power Theorem for Parametric, Longitudinal, Electron-Beam Amplifiers—H. A. Haus. (IRE Trans on Electron Devices, vol. ED-5, pp. 225–232; October, 1958. Abstract, Proc. IRE, vol. 47, p. 495; March, 1949.)

621.385.6:621.391.822 1067

New Method of Measuring the Noise Parameters of an Electron Beam—S. Saito. (IRE Trans. on Electron Devices, vol. ED-5, pp. 264–275; October, 1958. Abstract. Proc. IRE, vol. 47, p. 495; March, 1959.)

621.385.62 1068

A Multicavity Klystron with Double-Tuned

Output Circuit—H. J. Curnow and L. E. S. Mathias. (*Proc. IEE*, pt. B, vol. 106, pp. 487–488; September, 1959. Discussion, pp. 492–494.) An improved bandwidth (5.3 per cent–6.3 per cent for 2.5-mw output) has been obtained from a 6-cavity pulsed S-band klystron amplifier by the use of a double-tuned output circuit.

621.385.62:621.376.32 1069

A Simple Investigation of the Cross-Modulation Distortion arising from the Pulling Effect in a Freqency-Modulated Klystron—D. T. Gjessing. (*Proc. IEE*, pt. B, vol. 106, pp. 473–477; September, 1959. Discussion, pp. 492–494.) An expression for the distortion spectrum is derived theoretically, and families of curves showing severity of distortion are plotted. It is shown that small mis-matches of the output feeder give serious crosstalk.

621.385.624 1070

Large-Signal Analysis of the Multicavity Klystron—S. E. Webber. (IRE Trans. on Electron Devices, vol. ED-5, pp. 306–315; October, 1958. Abstract, Proc. IRE, vol. 47, p. 496; March, 1959.)

621.385.63 1071

Space and Time Harmonics in Electron Streams—R. Müller, (*Arch. elekt. Übertragung*, vol. 12, pp. 527–532; December, 1958.) Wave propagation is analyzed in systems with space-periodic structure which changes with time. These conditions arise when a high-intensity traveling wave interacts with an electron beam; harmonic frequencies are then generated. See also 2962 of 1957.

621.385.63 1072

General Design Procedure for High-Efficiency Travelling-Wave Amplifiers—J. E. Rowe and H. Sobol. (IRE Trans. on Electron Devices, vol. ED-5, pp. 288–300; October, 1958. Abstract, Proc. IRE, vol. 47, p. 495; March, 1959.)

621.385.63 1073

Design and Performance of Coupled-Helix Transducers for Travelling-Wave Tubes—T. S. Chen. (*J. Electronics Control*, vol. 6, pp. 289–306; April, 1959). Discussion of systems in which coupling between a traveling-wave-tube helix and the input or output coaxial cable is achieved by connecting the inner of the cable to a short helix wound round the outside of the main helix. The choice of pitch for the coupling helix is discussed and experimental results are given.

621.385.63 1074

Approximate Analytic Expressions for TWT Propagation Constants—W. H. Louisell. (IRE Trans on Electron Devices, vol. ED-5, pp. 257–259; October, 1958. Abstract. Proc. IRE, vol. 4, p. 495; March, 1959.)

621.385.63 1075

Travelling-Wave-Tube Propagation Constants for Finite Values of Gain per Wavelength—D. A. Dunn, G. S. Kino, and G. W. C. Mathers. (IRE Trans. on Electron Devices, vol. ED-5, pp. 243–251; October, 1958. Abstract, Proc. IRE, vol. 47, p. 495; March, 1959.)

621.385.63 1076

Improvement of Travelling-Wave-Tube Efficiency through Collector Potential Depression—F. Sterzer. (IRE Trans. on Electron Devices, vol. ED-5, pp. 300–305; October, 1958. Abstract. Proc. IRE, vol. 47, pp. 495–496; March, 1959.)

621.385.63 1077

Theory and Behaviour of Helix Structures for a High-Power Pulsed Travelling-Wave Tube—G. W. Buckely and J. Gunson. (*Proc. IEE*, pt. B, vol. 106, pp. 478–486; September, 1959. Discussion, pp. 492–494.) A multistart helix structure which may be water-cooled is considered theoretically and experimentally. There is reasonable agreement between measurements on 4- and 8-start structures and calculations of impedance and phase velocity on equivalent bounded sheath helices.

621.385.63 1078

Traveling-Wave-Tube Analogue—G. D. Sims and I. M. Stephenson. (*Electronic Tech.*, vol. 37, pp. 20–24, January, 1960.) The properties of any traveling-wave tube whose structure possesses equivalent parameters can be simulated, regardless of its actual physical form, by using the lumped-circuit principle and designing the inductance drift-tube structure to provide a given impedance and phase shift. The model described was designed primarily to investigate forward-wave interaction.

621.385.633 1079

Velocity and Current Distributions in the Spent Beam of the Backward-Wave Oscillator—J. W. Gewartowski. (IRE Trans. on Electron Devices, vol. ED-5, pp. 215–222; October, 1958. Abstract, Proc. IRE, vol. 47, pp. 494–495; March, 1959.)

621.385.633.1 1080

Influence of Magnetic Focusing Fields and Transverse Electron Motion on Starting Conditions for Spurious Oscillations in O-Type Backward-Wave Oscillators—H. G. Kosmahl. (IRE Trans. on Electron Devices, vol. ED-5, pp. 252–257; October, 1958, Abstract, Proc. IRE, vol. 47, p. 495; March, 1959.

621.385.69:621.372.2 1081

A New Wide-Band Thermionic Valve—T. Kojima. (*Onde élect.*, vol. 39, pp. 876–833; November, 1959.) Theoretical and practical considerations in the design of a distributed-amplifier-type tube are discussed [see 902 of 1954 (Lewis)]. Results obtained with a prototype tube operating from low frequencies up to 70 mc are compared with theoretical calculations.

621.387:621.362 1082

"Open-Circuit" Voltages in the Plasma Thermocouple—H. W. Lewis and J. R. Reitz. (*J. Appl. Phys.*, vol. 30, pp. 1838–1839; November, 1959.) Further discussion of the plasma diode (366 of January) clarifying the mechanism involved.

MISCELLANEOUS

621.38/.39:061.3 1083

Annual Convention of the Popov Society in Moscow, 12th–17th May, 1958—F. H. Lange. (*NachrTech.*, vol. 8, pp. 591–592; December, 1958.) A brief report on some of the papers presented at the convention, and a complete list of titles and authors covering the following four of twelve sections: a) information theory, b) antenna installations, c) semiconductor and miniature circuit elements, and d) electronics.

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ANE: Aeronautical and Navigational Electronics
AP: Antennas and Propagation
AU: Audio
AC: Automatic Control
BTR: Broadcast and Television Receivers
BC: Broadcasting (Formerly Broadcast Transmission Systems)
CT: Circuit Theory
CS: Communications Systems
CP: Component Parts
E: Education
ED: Electron Devices
EC: Electronic Computers
EM: Engineering Management
EWS: Engineering Writing and Speech
IE: Industrial Electronics
IT: Information Theory
I: Instrumentation
ME: Medical Electronics
MTT: Microwave Theory and Techniques
MIL: Military Electronics
NS: Nuclear Science
PT: Production Techniques
RFI: Radio Frequency Interference
RQC: Reliability and Quality Control
SET: Space Electronics and Telemetry (Formerly Telemetry and Remote Control)
UE: Ultrasonics Engineering
VC: Vehicular Communications

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Denison, A. B., Jr.: ME Dec 220, 277
Dennis, F. L.: CT Sep 323
Depian, L.: ED Apr 125
Deschamps, G. A.: AP Dec S371
Desmares, P.: BTR Dec 64
Desoer, C. A.: CT Sep 244, Dec 397
Deutsch, S.: CT Jun 214
Diamantides, N. D.: ANE Mar 16
Djorup, R. S.: SET Sep 148
Doherty, L. H.: AP Apr 117, Oct 419
Dolphin, L.: AP Apr 127
Dombrowski, G. E.: ED Oct 419
Dome, R. B.: AU Jan–Feb 16, Mar–Apr 29
Domeshek, S.: MIL Jul 87
Donald, D. D.: CS Jun 115
Dow, D. G.: ED Jan 106
Drubin, M.: CT Sep 325, Dec 386
Dubbe, R. F.: AU May–Jun 76
Duncan, J. W.: MTT Apr 257
Dunn, D. A.: ED Jul 294
Dunn, F. L.: SET Jun 82
Dunn, H. K.: AU May–Jun 47
Durkee, A. L.: CS Dec 272
Dwonczyk, M.: ANE Dec 228
Dyson, J. D.: AP Jan 102, Apr 181, Oct 329
Dzik, A.: VC Apr 58

E

Early, J. M.: ED Jul 322
Edwards, J. S.: ME Mar 39
Egan, J. F.: AC Dec 71
Ehrenpreis, D.: PT Jun 11; RQC Jun 70
Ehrenspeck, H. W.: AP Oct 379
Ehricke, K. A.: MIL Apr 58
Eisenberg, R. M.: MIL Jul 105
Elkind, J. I.: AC May 44
Ellerbruch, D. A.: CP Mar 42
Elliott, R. S.: AP Jan 87, Oct 435
Ellis, P. H.: AC Nov 43
Ellis, R. M.: ME Dec 204
Elspas, B.: CT Mar 45
Engen, G. F.: MTT Jul 351
Englund, J. W.: BTR Jan 13
Enslein, K.: I Mar 3, Sep 51; IE Jan 17; NS Mar 14
Enslow, G. M.: NS Jun 49
Erickson, K. E.: CT Mar 124
Espersen, G. A.: ED Jan 115
Evans, A. G.: AU Sep–Oct 116
Evans, D. H.: PT Jun 4
Evans, R. L.: ME Dec 266
Evans, W. R.: AC Dec 197

F

Fahey, W. J.: E Jun 75
Fannin, B. M.: AP Jul 258
Farber, R. J.: BTR May 50
Farmer, D. J.: MIL Oct 178
Farrall, W. R.: ME Dec 198
Feder, W.: ME Dec 240, 250
Feiner, A.: CS Dec 302
Feingold, W.: BTR Jan 10
Feinstein, J.: ED Jan 9
Feistman, M. L.: EWS Jan 22
Feldman, N. W.: VC Sep 27
Felsen, L. B.: AP Apr 168, Dec S231; MTT Jan 102
Ferguson, D. J.: ME Dec 291
Fettweis, A.: CT Sep 252
Fiedler, C. J.: AC Dec 81
Fielder, D. C.: CT Mar 81, Jun 158, Sep 282
Filipowsky, R.: CS Jun 95
Fink, B. R.: ME Sep 119
Fink, J. S.: ME Sep 141
Finkelstein, J.: PT Aug 50
Firle, T. E.: ED Jul 330
Fish, P. E.: BC Feb 14
Fisher, J. C.: EM Jun 46, Sep 65
Fiskin, J. M.: AP Jul 240
Fitch, M. E.: IE Jan 25
Fitzner, A. O.: IE Apr 19
Fleri, D.: MTT Jan 23
Flügge-Lotz, I.: AC Dec 109
Fogel, L. J.: ANE Dec 228
Foldes, P.: AP Jan 74
Fontaine, A. B.: CT May 60; IT May 60
Forgie, C. D.: AC May 44
Fosberg, A. M.: ME Dec 246
Foster, J. H.: RFI May 11
Fountain, A. M.: EWS Jun 46
Fournier, R. V.: BTR Dec 46
Fowler, A. D.: CS Sep 173
Fowler, E.: AU Nov–Dec 169
Fraade, D. J.: IE Dec 1
Frank, K.: ME Jun 85
Frankel, S. P.: EC Sep 271
Franklin, D. L.: ME Dec 204
Franz, K.: AP Oct 353, Dec S68
Freethey, F. E.: MTT Jul 388
Freidland, B.: CT Mar 61; IT Sep 114, 136
Friedman, B.: AP Dec S227
Freidman, M. P.: ME Sep 121
Freiman, C. V.: CT May 71; IT May 71, Jun 62
Freimer, M.: AC Nov 10; IT Mar 25
Frihart, H. N.: BTR Dec 54
Froelich, R. W.: AC Dec 139
Frost, H. B.: ED Jul 315
Frumin, M. J.: ME Mar 30
Fry, D. L.: ME Dec 252, 259, 264
Fryer, W. D.: CT Jun 144
Fujisawa, T.: CT Sep 261
Fukada, M.: CT Sep 277
Fullwood, R.: EC Dec 498
Furutsu, K.: AP Dec S209

G

Gabriel, W. F.: MTT Oct 481, 482
Galbraith, R. A.: E Apr 50
Galejs, J.: IT Jun 79, Sep 106; SET Dec 179
Gallagher, H. E.: ED Oct 390
Gamo, H.: CT May 283; IT May 283
Gardner, C. S.: AP Dec S87
Gariano, W. F.: ED Jan 90
Garner, H. L.: EC Mar 25, Jun 140
Garver, R. V.: MTT Apr 272
Gates, C. R.: AC Dec 34
Gau, L.: IE Jul 43
Geddy, G. B.: PT Aug 27
Geffe, P. R.: CT Mar 129, Dec 395
Gerardi, F. R.: CT Jun 197
Gerhold, R. A.: PT Aug 12
Germano, C. P.: AU Jul–Aug 96
Gerosa, G.: AP Dec S471
Gerst, I.: MTT Oct 475
Getting, I. A.: MIL Jan 19
Gianoplus, A. S.: SET Jun 66
Giesselman, A. C.: VC Sep 4
Giguere, W. J.: EC Dec 432
Gilbert, E. G.: EC Jun 204
Gilbert, E. O.: EC Sep 391
Gill, A.: EC Dec 486; IT Jun 52
Gillespie, E. F. F.: AP Dec S219
Ginsburg, S.: EC Mar 13, Sep 346, Dec 441
Ginzton, E. L.: ED Jan 64
Glass, H. I.: MTT Apr 295
Glenn, A. B.: CS May 14
Glomb, W. L.: CS Dec 284
Golay, M. J. E.: SET Dec 186
Gold, B.: IT Mar 17, 25
Gold, L.: SET Dec 162
Gold, R. D.: EC Sep 287
Goldin, P. J.: PQC Jun 7
Goldman, I.: EM Jun 55
Goldstein, H.: ANE Jun 142
Goldstone, L. O.: AP Jul 274, Oct 307; MTT Apr 213
Golomb, S. W.: CT May 176; IT May 176
Gonzales, C.: VC Sep 83
Goodrich, R. F.: AP Jul 213, Dec S23
Gordon, E. S.: BTR Dec 8
Gordon, J.: VC Apr 74
Gorozdos, R. E.: AC Dec 46
Goubau, G.: AP Dec S140
Goudet, G.: MTT Jul 327
Graham, M.: EC Dec 479
Graham, R. H.: NS Dec 16
Grasberg, A. G.: E Jun 94; EM Mar 31
Gray, C. L.: AP Jul 281
Green, H. D.: ME Dec 277

Green, J. H.: VC Apr 74
Green, M.: CS May 47
Greenleaf, R. K.: EM Mar 19
Greifinger, C.: ED Jul 288
Grimm, H. J.: I Dec 97
Grinich, V. H.: EC Jun 108
Groginsky, H. L.: ANE Sep 111
Grosswald, E.: CT Dec 334
Guilford, E. C.: E Sep 138
Guillemin, E. A.: CT May 110, IT May 110

H

Haas, I.: EC Jun 108
Hadady, R. E.: SET Mar 14
Haddad, R. A.: AC Nov 121
Hadley, A. M.: CP Sep 175
Haefeli, H. G.: CT May 79; IT May 79
Hall, N. A.: E Apr 43
Halstenberg, R. V.: AC Dec 97
Hamasaki, J.: MTT Jul 360
Hamburger, F., Jr.: E Apr 31
Hammerle, K. J.: IT Mar 9
Handelsman, M.: CS May 31
Hanley, G.: MTT Jan 23
Hansen, R. C.: AP Jul 276, Dec S458
Harary, F.: CT May 59; IT May 95
Hare, M. D.: ED Oct 397
Harken, D. E.: ME Dec 246
Harrington, R. F.: AP Apr 150
Harris, C. M.: AU May–Jun 80
Harris, E. F.: AP Jul 281; VC Apr 2
Harris, L. A.: ED Oct 413
Hartel, R. R.: EC Jun 159
Hartin, W. J.: NS Mar 11
Hartline, H. K.: ME Jun 84
Hartmanis, J.: CT Mar 69
Harvey, A. F.: MTT Oct 402
Harvey, R. A.: NS Dec 20
Hassler, G. L.: ME Mar 52
Hauser, G.: RQC Jun 53
Hawkins, H. E.: ANE Mar 26
Haycock, O. C.: AP Oct 414
Hayes, O. E.: ED Jul 330
Hazel, H. K.: PT Aug 39
Hazony, D.: AC Nov 132
Heald, M. A.: NS Sep 33
Heckelman, T. J.: CS Jun 142
Hedgecock, N. E.: AP Dec S284
Hedrich, A. L.: I Mar 22
Heffner, H.: MTT Jan 83
Heider, E.: AP Apr 201
Held, G.: AP Dec S296
Hellman, R. K.: E Sep 113
Hellstrom, M. J.: BTR May 7
Helstrom, C. W.: IT Sep 139
Helvey, T. C.: SET Jun 57, Sep 99
Henn, W.: AC Dec 116
Hennie, F. C.: CT Mar 35
Henny, G.: ME Jun 75
Herold, E. W.: NE Sep 1
Herrick, J. F.: ME Dec 195, 202
Herrman, G. F.: BTR Dec 75
Herrold, G.: ME Dec 213
Hessel, A.: AP Dec S201
Hewish, A.: AP Dec S120
Heyne, J. B.: RQC Jun 41
Hicks, B. L.: CT Mar 137
Higgins, S. F.: SET Sep 123
Higgins, T. J.: AC Dec 150; ME Sep 125
Hilibrand, J.: EC Sep 287
Hill, R. M.: MTT Jan 73
Hillyard, L. R.: E Mar 10
Hoadley, G. B.: E Sep 122
Hobbs, E. W.: CT Mar 135
Hoch, O. L.: ED Jan 18
Hodder, W. K.: SET Sep 117
Hoehndorf, F.: MIL Oct 162
Hoffman, R. A.: MIL Oct 160
Hoffman, W. C.: AP Dec S301
Hoge, R. R.: IE Jan 38; NS Jun 42
Holland, D. B.: ME Sep 125
Holland, G.: CS Sep 180
Hollis, J. L.: CS Sep 185
Holt, S. F.: AP Oct 434
Honey, R. C.: AP Oct 320
Hoover, G. W.: PT Aug 23
Horowitz, I. M.: AC Dec 5; CT Sep 296
Horsting, C. W.: ED Jan 119
Hosono, T.: MTT Jul 370
Hougardy, R. W.: AP Jul 276
Houseman, E. O., Jr.: MTT Oct 475
Howarth, D. J.: CP Jun 81
Hoyt, W. M.: E Mar 10
Hsieh, H. C.: AC Nov 16, Dec 212
Hu, M. K.: AP Oct 373
Hu, Y. Y.: AP Oct 373
Huang, C.: ED Apr 141, 154
Huber, J. C., Jr.: AP Jul 281
Hudson, A. C.: MTT Apr 277
Hudson, W. T.: E Mar 10
Huffman, D. A.: CT Mar 4, May 41; IT May 41
Hufnagel, C. A.: ME Mar 13
Huggins, W. H.: AC Nov 91
Hughes, H. E.: ED Jul 311
Hull, J. L.: PT Aug 50
Hupert, J. J.: CT Sep 292
Hyneman, R. F.: AP Oct 335

I

Iden, F. W.: ANE Jun 100
Igleheart, J. C.: CS Sep 173
Iizuka, K.: AP Oct 386
Ikegami, F.: AP Jul 252
Ikeno, N.: CT May 187; IT May 187
Ingram, S. B.: E Apr 58
Inouye, G. T.: MIL Oct 178
Isley, C. T.: IT Sep 139, Dec 186
Ivey, H. F.: ED Apr 203, Jul 335

J

Jackson, H. G.: NS Jun 64
Jackson, W. E.: ANE June 85
Jacobsen, A. B.: PT Aug 28
Jacobson, B. S.: ME Jun 66
Jacobson, E. D.: ME Sep 116
Jacobson, M. J.: IT Mar 4
Jameson, W. P.: MIL Jul 105
Jamison, J. H.: EC Dec 432
Janos, W. A.: IT Jun 67
Jaynes, E. T.: IT Sep 98
Jewett, R. H.: MIL Jan 12
Jewett, W. M.: CS May 7
Jochim, K. E.: ME Dec 232
Johler, J. R.: AP Jan 1
Johnson, C. C.: ED Jan 6, Apr 189, Oct 409
Johnson, C. M.: MTT Jan 27, Apr 229
Johnson, C. W.: ME Sep 134
Johnson, L. J.: CP Dec 259
Johnson, P.: ME Dec 287
Johnson, R. C.: MTT Jul 374
Johnston, C.: VC Sep 24
Johnston, L. B.: BTR May 54
Johnston, T. W.: AP Dec S337
Jones, E. M. T.: MTT Jan 128, Oct 453
Jones, H. C.: RQC Jun 16
Jones, R. E.: ME Jun 92
Jordan, E. C.: AP Jan 87
Jordan, W. F., Jr.: ED Jan 79
Junga, F. A.: NS Jun 49
Jury, E. I.: AC May 15

K

Kahn, L.: CP Sep 150
Kahn, W. K.: MTT Jan 102, Oct 430
Kalaba, R.: AC Nov 1; CT May 271; IT May 271
Kalman, R. E.: AC Dec 20, 110, 112
Kamenetsky, M.: MIL Jul 75
Kane, W. E.: RQC Jun 34
Kantrowitz, A.: MIL Apr 34
Kaplan, L. J.: MTT Apr 298, Jul 392, Oct. 473, 475
Karal, F. C., Jr.: AP Dec S91
Karbowiak, A. E.: AP Dec S191
Karp, S. N.: AP Dec S91
Karr, P. R.: IT Mar 33, Sep 140, Dec 186
Katsura, S.: ME Dec 283
Katz, A.: EM Sep 75
Kawahashi, T.: MTT Apr 247
Kay, A. F.: AP Jan 22, 32
Kay, I.: AP Jan 11, Dec S255
Kazarinoff, N. D.: AP Dec S21
Kear, F. G.: ANE Jun 61
Kearney, J. W.: E Jun 67
Keen, A. W.: CT Jun 188
Keilson, J.: IT Jun 75, 77
Keller, E. A.: IE Apr 35
Keller, J. B.: AP Apr 146, Dec S52, S87
Kelley, L. C.: CS Sep 167
Kenna, L. A.: EM Jun 49
Kennedy, W, F.: SET Sep 138
Kent, G.: CT Dec 388
Kenyon, R. J.: MTT Jul 391
Kessler, E., III: ANE Mar 31
Keys, J. E.: AP Dec S77
Kie , K. M.: CS Jun 125
Kilmer, W. L.: E Jun 85; EC Sep 321
Kim, C. S.: BTR Dec 83
Kim, W. H.: CT May 71, Jun 219, Sep 267; IT Jun 62; AC Dec 108
King, D. D.: MTT Apr 229
King, E. N.: PT Jun 18
King, H. E.: AP Apr 197
King, R.: AP Jan 53, Dec S440
Kingston, R. H.: MTT Jan 92
Kintner, P. M.: EC Jun 227
Kirkpatrick, J.: ME Jun 82
Kirschbaum, H. S.: MTT Jan 142
Klante, K.: AP Dec S68
Klapper, J.: AU May–Jun 80
Klein, J. J.: AC Nov 182
Klein, R. J.: VC Apr 13
Kleinman, R. E.: AP Jul 213
Kline, D. F.: E Apr 38, Jun 108
Klipsch, P. W.: AU Mar–Apr 34, Jul–Aug 93, Nov–Dec 161
Knight, T. G.: CS Sep 209
Knudsen, H. L.: AP Dec S361
Knudsen, H. K.: AC Nov 112
Ko, W.: ED Jul 341
Koch, J. W.: CS Jun 77
Kochenburger, R. J.: AC Dec 59
Kodis, R. D.: AP Dec S468
Koenig, H. E.: E Sep 128
Kogo, H.: MTT Jul 393
Kohno, S.: MTT Jul 370
Kolff, W. J.: ME Mar 7
Kolodkin, S. S.: SET Mar 47
Konrad, M.: NS Mar 35
Kosmahl, H. G.: ED Apr 225
Kosonocky, W. F.: EC Sep 277
Kouyoumjian, R. G.: AP Dec S273, S379, S406
Kranc, G. M.: AC Dec 108
Krassner, G. N.: CS Dec 232
Krendel, E. S.: ME Sep 149
Kroll, N.: EC Sep 307
Kudravcev, V.: ME Dec 267
Kuether, F. W.: ME Dec 286
Kummer, W. H.: AP Oct 428
Kunze, A. A.: CS Dec 241
Kuo, F. F.: CT Mar 131
Kurokawa, K.: MTT Jul 360
Kurshan, J.: EM Mar 14

L

Ladd, A. W.: CS Jun 102
Ladd, D. W.: EC Mar 36
Lai, D. C.: AC Nov 91
Lakin, J. R.: E Apr 50
Lambert, L. M.: EC Dec 470
Lambert, V. L.: EM Mar 8
Landy, J. J.: AC Dec 81
Lane, J. A.: MTT Jan 177
Lane, W. V.: PT Aug 12
Langberg, E.: CS Jun 68
Lange, A. S.: AC May 21
Lanzkron, R. W.: AC Dec 150
La Plant, O.: ANE Mar 26
La Rosa, R.: CT Mar 95
Lass, H.: IT Sep 138
Latimer, K. E.: CP Jun 62
Leadabrand, R. L.: AP Apr 127
Lebenbaum, M. T.: E Jun 67
Ledley, R. S.: EC Jun 131, 230
Lee, H. E.: BTR May 25
Lee, L. K.: PT Jun 22
Lee, R. C.: EC Jun 186
Leipnik, R.: IT Dec 184
Leith, E. N.: AC Nov 137
Leondes, C. T.: AC Nov 16, 100, Dec 212
Lerner, R. M.: CT May 197; CP Dec 263; IT May 197
Lesk, I. A.: ED Jan 28
Levadi, V. S.: AC Nov 55
Levey, L.: MTT Jan 102
Levin, E.: AP Apr 142
Levine, A.: IT Jun 90
Levine, D.: AC May 69
Levine, M. B.: IE Dec 26
Levis, C. A.: AP Dec S424
Levy, B. R.: AP Dec S52
Levy, E. C.: AC May 37
Lewin, L.: AP Apr 162
Lewis, C. H.: CP Sep 144
Lewis, D. J.: MTT Jan 110
Li, T.: AP Jul 223
Li, Y. T.: IE Dec 57
Lin, G.: ME Dec 277
Lin, H. C.: CP Dec 269; ED Jan 79
Linden, D. A.: CT Sep 323
Lindorff, D. P.: AC Dec 173
Lineweaver, J. L.: ED Jan 118
Litchford, G. B.: ANE Jun 118
Lo, A. W.: EC Sep 277
Lock, J. M.: CP Jun 93
Lofgren, L.: CT May 158; IT May 158
Loh, S. C.: AP Jan 46, Jul 226
London, F. H.: IE Jan 31
Lory, H. J.: AC Nov 91
Losee, F. A.: CS Jun 102
Lucal, H. M.: EC Dec 449
Luebke, W. R.: ED Jul 294
Luedicke, E.: BTR May 33
Lukatela, G. L.: IE Jul 40
Lunelli, L.: CT Dec 392
Lutz, S. G.: CS Jun 102

M

Macdonald, J. R.: AU Sep–Oct 128
Mack, D. A.: NS Jun 64
Mackay, R. S.: ME Jun 100
Maclean, T. S. M.: AP Dec S379
MacPherson, A. C.: ED Jan 83; I Sep 44
MacWilliams, F. J.: CT Mar 135
Madden, R.: ANE Dec 219
Madich, P.: EC Jun 182
Madrid, D. C.: SET Jun 73
Maffett, A. L.: AP Jul 213
Mager, R. F.: E Jun 104
Maher, T. M.: AP Apr 199
Makow, D. M.: ANE Sep 222; I Mar 32
Mangiaracina, R. S.: MTT Jan 11
Manke, A. G.: VC Apr 47
Mann, P. A.: AP Oct 353
Marcus, M. B.: IT Dec 179
Marcus, R. S.: IT Mar 12
Marcus, S. M.: EC Jun 98
Marden, E.: ME Sep 167
Margolis, M.: AC Nov 100
Marie, P.: AP Dec S412
Markham, A. S.: ANE Mar 4
Martin, A. V. J.: CS May 57
Martin, M. A.: SET Mar 33
Martinek, J.: ME Sep 112
Martinengo, R.: BTR Dec 31
Maslen, S. H.: MIL Apr 52
Mason, H. L.: ME Sep 167
Mathews, M. V.: IT Sep 129
Matsumaru, K.: MTT Jan 175, Apr 192
Matthaei, G.: MTT Oct 453
Matthews, L. E.: BTR Jan 22
Matzek, T.: BTR Jan 18
Mayeda, W.: CT Mar 136, Dec 394
McAuliffe, G. K.: CS Sep 189
McClurg, G. O.: IE Dec 20
McCluskey, E. J., Jr.: EC Dec 439
McConnell, D.: ME Sep 121
McCook, R. D.: ME Dec 234
McCormick, G. C.: AP Dec S288
McFadden, J. A.: CT May 228; IT May 228, Dec 174
McFarlan, R. L.: E Mar 23
McFee, R.: AP Apr 199
McGhee, R. B.: AC Dec 199; IT Jun 90
McGregor, J. E.: PT Sep 30
McGregor, R. R.: ME Mar 51
McKain, J. L.: AU Jul–Aug 101
McLay, A. B.: AP Dec S284
McLean, J. D.: EM Sep 71
McMillan, S. H.: IE Jan 34
McPhee, K. H.: CP Mar 3
McRae, D. D.: SET Jun 61
Meagher, R. E.: EC Sep 263
Medd, W. J.: MTT Jul 395
Meixner, J.: AP Dec S435
Melchor, J. L.: MTT Jan 15
Melmed, A.: EC Dec 474
Melngailis, I.: RFI May 11
Mermin, N. D.: IT Jun 75, 77
Metcalfe, D.: CS Dec 272

Metz, H. I.: ANE Jun 78
Metze, G.: CT Mar 5
Meyer, D. R.: ME Sep 121
Michaelson, H. B.: EWS Dec 89
Mickelson, N. B.: MIL Jul 69
Mickelwait, A. B.: MIL Oct 149
Middleton, D.: CT May 234; IT May 234, Jun 86
Mihran, T. G.: ED Jan 54
Mikelson, W.: AC Dec 21
Miller, F. B.: IE Jul 3
Miller, G. L.: EC Dec 479
Miller, N. D.: CS Dec 249
Miller, R.: BTR Jan 5
Miller, R. E.: CT Mar 5
Milligan, J. V.: ME Dec 274
Milnes, A. G.: E Mar 6; ED Apr 125; EC Dec 458
Mine, H.: CT May 138; IT May 138
Mishkin, E.: AC Nov 121
Mitchell, J. F.: VC Sep 20
Mittra, R.: AP Dec S244
Miyata, J. J.: EC Jun 159
Moglia, G. A.: PT Jun 29
Montgomery, L. H.: ME Mar 29
Montgomery, P. O'B.: ME Sep 186
Montgomery, R. A.: MIL Jan 12
Mood, A.: RQC Jun 16
Moore, G. E.: E Jun 82
Moore, R.: ME Sep 171
Morgan, S. P.: AP Oct 342
Morgan, W. L.: EC Jun 148
Morgan-Voyce, A. M.: CT Sep 321
Morgenthaler, F. R.: MTT Jan 6
Morris, R. M.: BC Feb 22
Morrison, J. A.: ED Apr 231
Morrison, R.: CT Sep 310
Mortenson, K. E.: ED Apr 174
Muchmore, R. B.: AP Apr 142
Muehe, C. E.: MTT Apr 296
Mueller, C. W.: EC Sep 287
Mueller, G. E.: SET Dec 196
Muldrew, D. B.: AP Oct 393
Muller, D. E.: EC Sep 401
Mulligan, J. H., Jr.: EC Mar 48
Mullin, A. A.: CT Mar 131; EC Dec 498, 499; IT Sep 137
Mullin, F. J.: AC May 15
Muniz, R.: BTR Jan 76
Munson, J. K.: EC Jun 200
Murakami, T.: BTR Jan 46
Muroga, S.: EC Sep 308
Murphy, G. J.: AC Dec 71
Murphy, W. A.: EWS Dec 75
Myrick, J. C.: CS Sep 180

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Nakada, T.: AC Dec 37
Nakahara, T.: MTT Jul 366
Nalos, E. J.: ED Jan 48
Nash, C. W.: ME Dec 274
Navarro, J. A.: RQC Jun 64
Navon, D.: ED Apr 169
Neal, G.: AP Apr 117
Nelson, C. V.: ME Sep 107
Nelson, J. G.: PT Aug 51
Nelson, R. E.: ED Jul 270
Nelson, R. J.: VC Sep 83
Nelson, S. W.: PT Sep 1
Nesh, F.: UE Feb 3
Netherwood, D. B.: EC Sep 339, 367
Neugebauer, H. J.: AP Dec S226
Neuringer, J. L.: SET Jun 55
Neville, W. E.: ME Mar 50
Nevins, J. E.: ED Apr 195
Newell, W. E.: ED Apr 125
Nicholas, J. C.: PT Aug 28
Nicholson, N. N.: EWS Jun 52
Niehaus, D. J.: IE Jan 38; NS Jun 42
Nishino, H. H.: EC Sep 326
Noble, B.: AP Dec S37
Noble, H. V.: CP Sep 210
Noggle, L. H.: E Jun 78
Noll, J. C.: EC Dec 432
Nordenberg, H. M.: CP Sep 201
Northover, F. H.: AP Dec S340
Novak, J. F.: AU Jan–Feb 5
Nucci, E. J.: CP Sep 128

O

Och, H. G.: MIL Jan 8
Ochs, G. R.: CS Dec 263
O'Hara, F. J.: MTT Jan 32
Oldenburger, R.: AC May 5, Dec 37
Oliner, A. A.: AP Apr 153, Jul 274, Oct 307, Dec S201; MTT Jan 134
Olmsted, F.: ME Dec 210
Olte, A.: AP Dec S61
O'Meara, T. R.: CP Jun 49, Dec 273
O'Neal, C. D.: AU Nov–Dec 140
Onyshkevych, L. S.: EC Sep 277
Orchard, H. J.: CT Dec 370
Orlando, P.: EC Jun 218
Ornstein, W.: VC Sep 64
Ortel, W. C. G.: EC Sep 265
Otsuka, S. P.: ED Jan 48
Otterman, J.: I Mar 8, Sep 55
Overstreet, R. L.: I Sep 46
Owen J.: MTT Jul 384

P

Packer, L.: MIL Jul 114
Palmer, J. L.: ED Jul 262
Pantell, R. H.: ED Jan 48
Papadopoulos, V. M.: AP Dec S78
Papoulis, A.: AC Nov 67; CT Mar 129
Parezanovich, N.: EC Jun 182
Parker, H. N.: BTR Dec 70
Parker, N. W.: BTR Dec 54
Parry, C. A.: CS Sep 211, Dec 290
Pastel, M. P.: AC Nov 192
Patapoff, H.: AC Dec 139
Pate, H. R.: EC Dec 479
Patrick, D.: ME Dec 287
Patterson, K. G.: MTT Oct 466
Paull, M. C.: EC Sep 356; IT Mar 34
Paulsen, R. C.: EC Jun 169
Peake, W. H.: AP Dec S324
Pearlman, L. S.: VC Apr 54
Peirce, E. C., II: ME Mar 54
Peiss, C. N.: ME Dec 234
Peless, Y.: RFI May 18
Penndorf, R.: CS Jun 121
Perkins, J. F., Jr.: ME Mar 24
Perlman, S.: CS Sep 167
Perlmutter, A.: CP Sep 180
Peterson, A. M.: AP Apr 127
Peterson, A. P. G.: EWS Dec 73
Peterson, G. E.: AU Nov–Dec 148
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  Wedge and Septate, Attenuation in: AP Dec S279
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X-Guide and *O*-Guide Surface Wave Transmission: MTT Jul 366

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Sample cell used at the IBM Watson Laboratory for studies of the self-diffusion of quantum liquids.

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X	MA-710	8.5-9.6 kMc	30 db min.	0.2 db max.	1.10 max.
X	MA-750*	8.5-9.6 kMc	30 db min.	0.2 db max.	1.10 max.
Ku	MA-760	16.0-17.0 kMc	30 db min.	0.2 db max.	1.10 max.
Ku	MA-776**	16.0-17.0 kMc	75 db min.	0.2 db max.	1.10 max.
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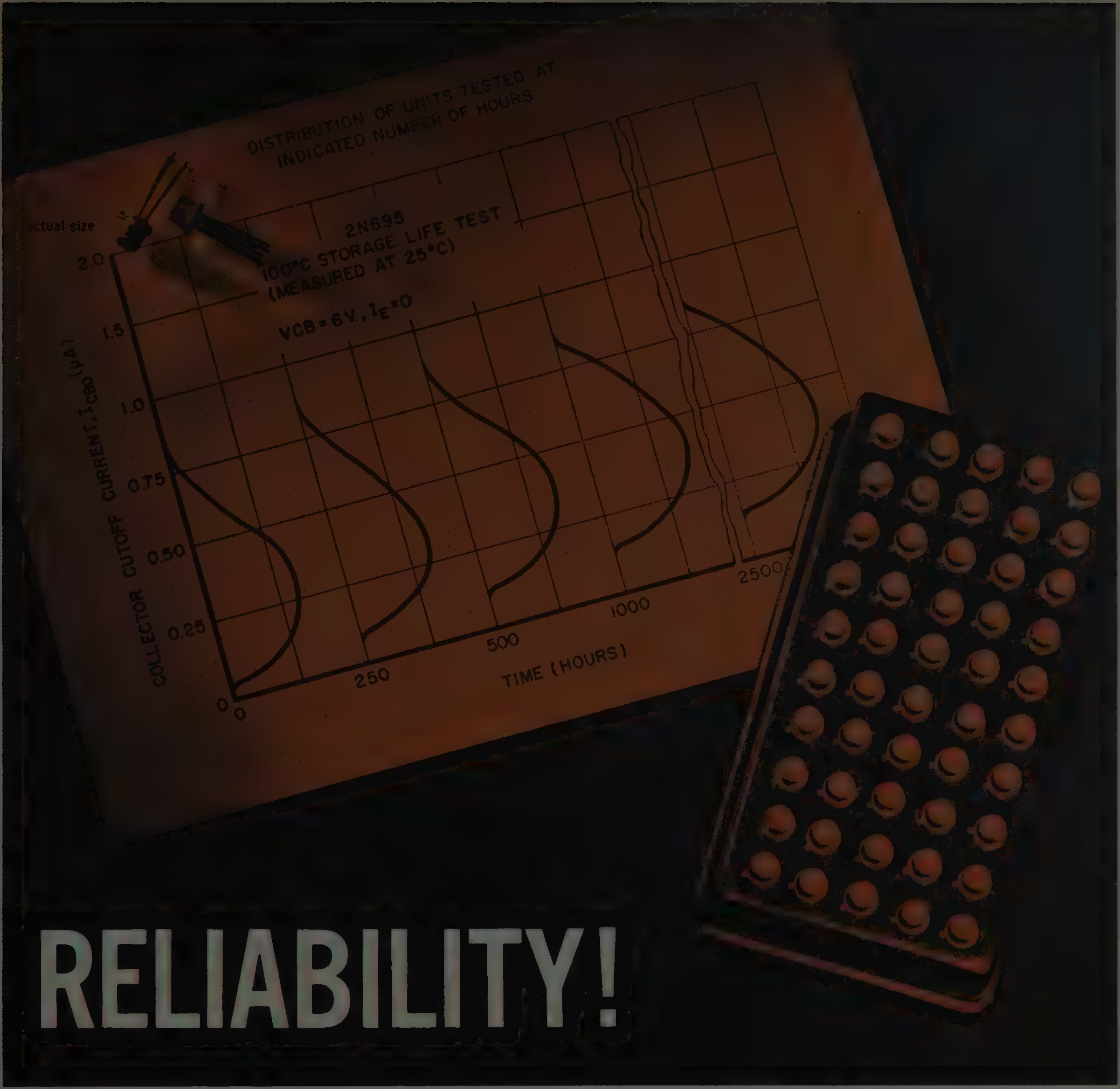

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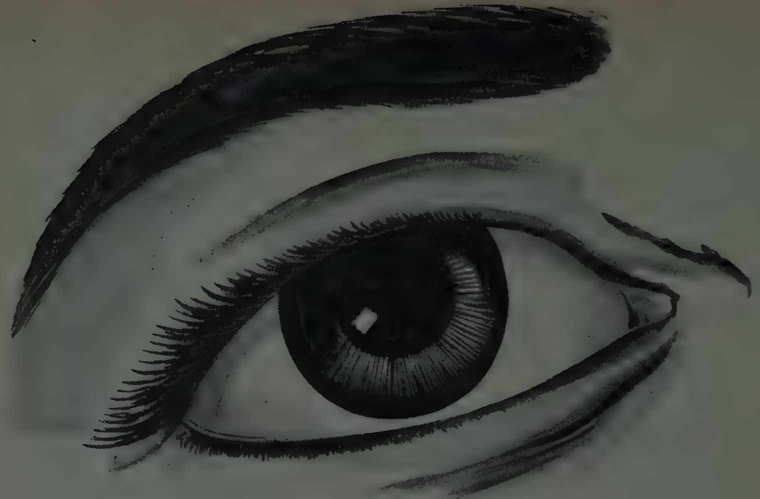

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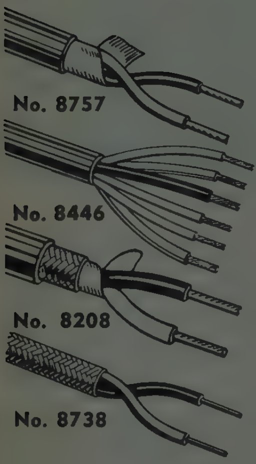

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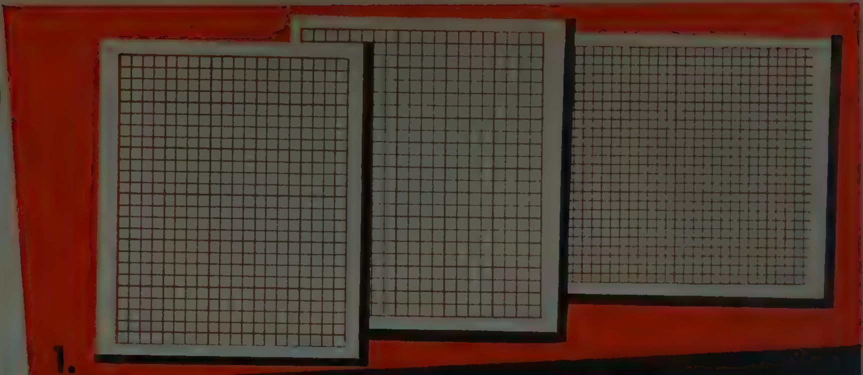
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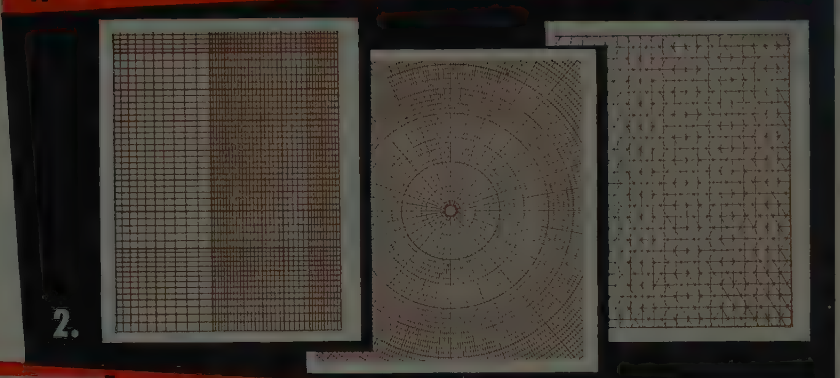
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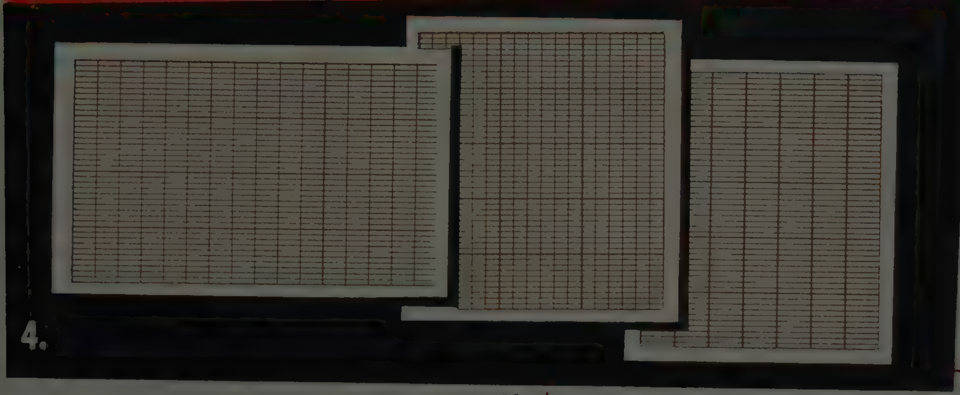
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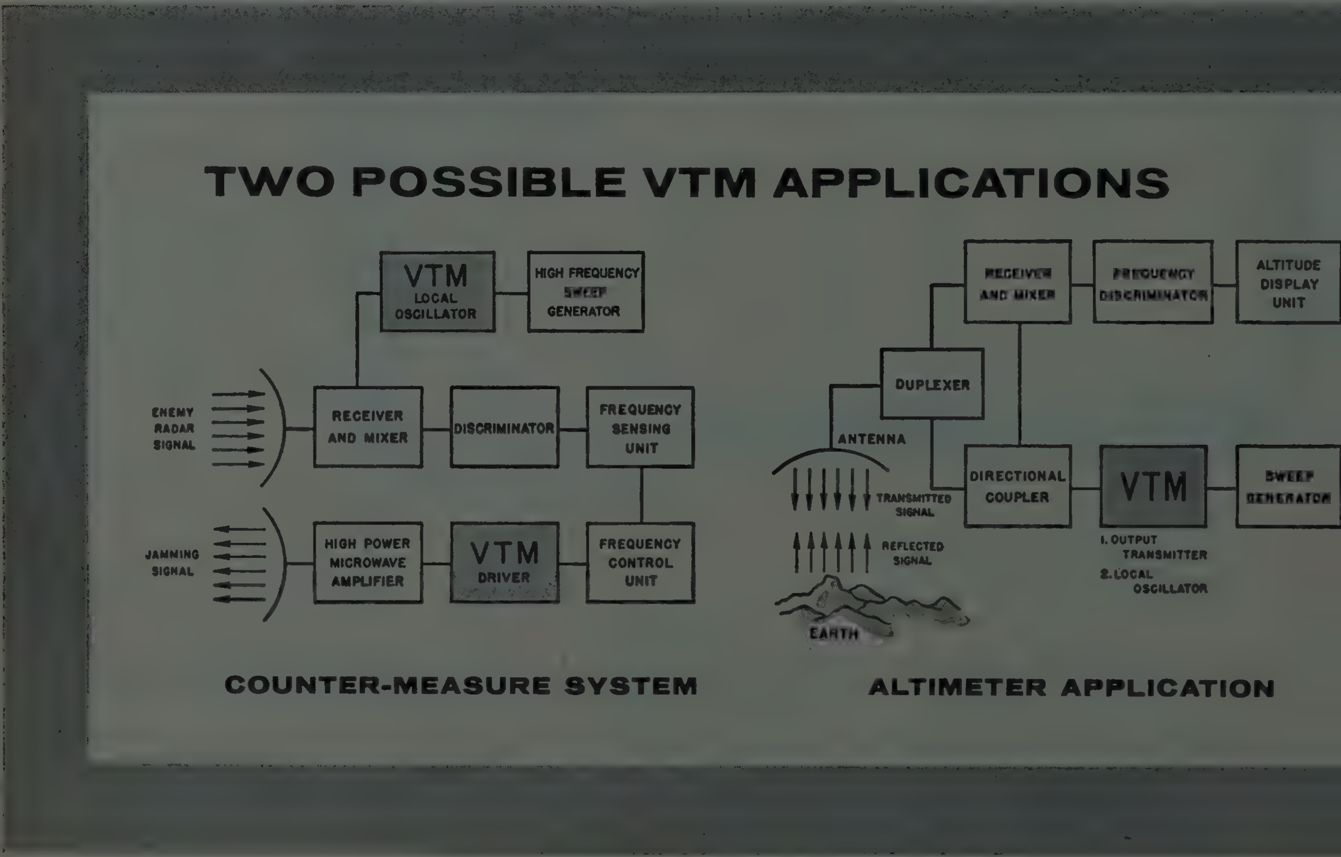
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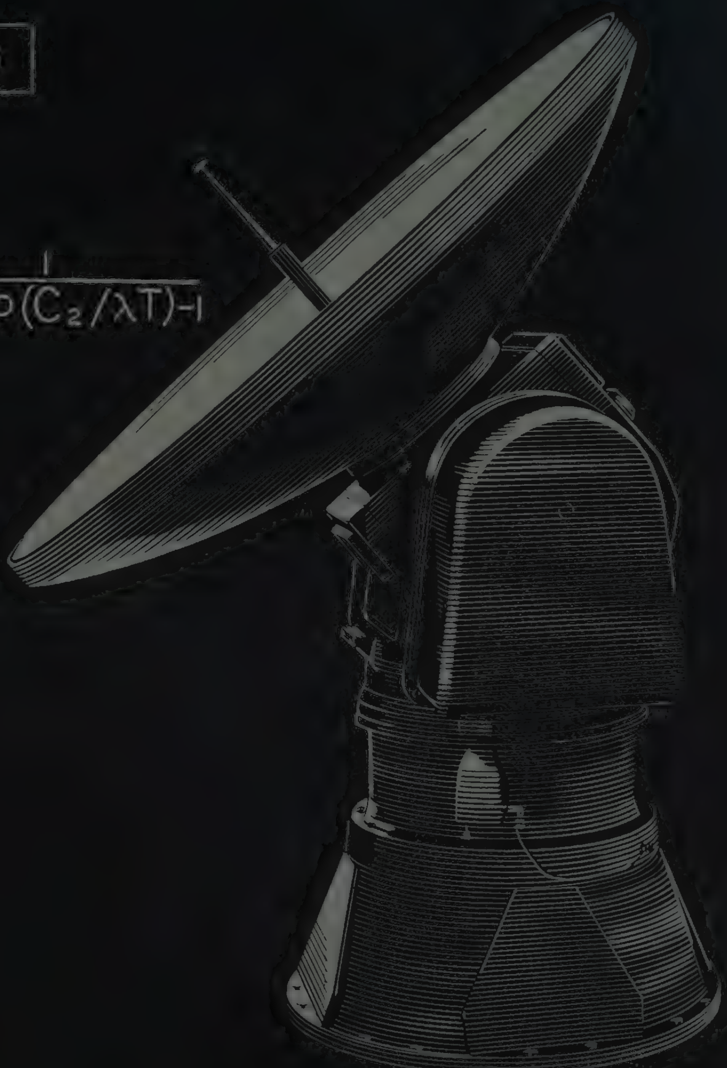
$$C_1 = 8\pi hc$$

$$C_2 = \frac{hc}{k}$$



THE UNUSUAL SHAPE OF PROGRESS

The principles of spectral energy distribution of the radiation from a black body, announced by Max Planck in 1900, were of intense interest to the theoretical physicists of the day. Since then, these principles have paved the way for major advances in marine navigation and radio astronomy. For example, Collins new Radio Sextant pictured here is capable of continuously tracking the sun or the moon under any weather conditions, furnishing both the ship's location and heading with high precision.

Max Planck's work was done in an atmosphere of unrestricted scientific freedom. Such an atmosphere is provided in Collins Radio Company Research laboratories for physicists, mathematicians and engineers engaged in basic research. These are men capable of looking beyond man's present limitations, with the ability and ambition to analyze man's progress and envision his future environment. To further implement the advancement of scientific knowledge at Collins, with the resultant development of new technologies, unique professional opportunities are now being offered in the fields of radio astronomy, circuits, advanced systems, antennas and propagation, mechanical sciences and mathematics. Your inquiry is invited.

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- ☐ Microwave
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- ☐ Antenna
- ☐ Servo-Mechanisms
- ☐ Navigation
- ☐ Counter Measures
- ☐ Telemetering
- ☐ Nucleonics
- ☐ Ind'l. Instruments
- ☐ Components
- ☐ Circuit Design

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Positions Open

(Continued from page 116A)

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Electronics Research and Development Lab. needs man with experience in transistor-circuit development. Position carries opportunity to study for advanced degrees in Physics or Engineering. Salary consistent with training and experience. Applicants will be required to have BS. degree or better in Physics or Engineering to qualify. For more details write Research Foundation, Oklahoma State University, Stillwater, Okla., or call R. F. Buck, FRontier 2-6211, Ext. 579, after 5 P.M. or holidays call FRontier 2-8480.

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Expert in field of semiconductors. Ph.D. is minimum requirement, applicant must have some teaching experience. Academic rank and salary dependent on demonstrated ability and over-all experience. The man in this position will divide his time between teaching and research. Appointment begins Sept. 1960. Write, including resume to W. R. Beam, Head, E.E. Dept., Rensselaer Polytechnic Institute, Troy, N. Y.

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Scientific book publisher needs radio engineer to plan and develop new books. Publishing experience desirable. Write in detail. State salary. Box 2012.

ENGINEERS

Major firm seeking (1) Section leader, Electromechanical design, (2) Section leader, testing and reliability, knowledge of Military acceptance testing, (3) Manufacturing engineers, knowledge of standards, process and tool engineering for electromechanical display equipment. Top future positions. Benefit programs. Salaries $8,000–$15,000. Contact Mr. W. Hart, Witty-Polon Management Consultants, 176 East 75 St., New York 21, N. Y. Telephone: Lehigh 5-4222.

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For those interested in teaching fundamental electrical & electronic subjects. Opening for fall semester beginning Sept. 1960. Salary depends

(Continued on page 120A)

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DIGITAL COMPUTER DESIGN
ADVANCED ANTENNA SYSTEMS
ADVANCED SYSTEMS AUTOMATED TEST
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To join a development team on high and low powered Traveling Wave Tubes and Carcinotrons.

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MAGNETRON
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Openings exist in these product line areas for Junior, Intermediate, and Senior level engineers.

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If you qualify for any of these positions, please contact Mr. Thomas L. Fike, 960 Industrial Road, San Carlos, California.

LITTON INDUSTRIES
Electron Tube Division
San Carlos, California

Positions Open

(Continued from page 118A)

on qualifications. New building and equipment. Opportunity for advancement and further study. 9 month academic year. Part time evening session work available. Health Insurance and Pension Plan. Send complete resume to Prof. I. L. Kosow, Dept. Head. Electrical Technology, Staten Island Community College, 50 Bay St., Staten Island 1, N. Y.

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Fast growing Engineering School on west coast needs 3 electronic-electrical engineers able to teach circuits and electronics, with a specialty in communications, industrial electronics or computers. MS. degree plus industrial experience is the minimum requirement; a Ph.D. is desirable. Salary and rank will depend on experience. The dept. has just moved into a new building and needs aggressive men who wish to develop curricula and labs. Apply L. Cromwell, Head, E.E. Dept., Los Angeles State College, 5151 State College Dr., Los Angeles 32, Calif.

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Biophysical Electronics, Inc. devotes its efforts to the design and manufacture of instruments assisting the physician, psychologist, and research scientist to meet the challenge of preserving and extending life. Positions are open for: Senior Transistor Circuit Designer, Sales and Office Manager and experienced Electronic Technician. Submit resume, availability and salary requirements in first letter to Biophysical Electronics, Inc., New Hope, Pa.

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(Continued on page 122A)

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DEVICE DEVELOPMENT ENGINEER. *Mission:* Probing the unknown to evolve a variety of new advanced semiconductor-component devices.

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Device and Circuit Development Engineers we seek, want liberal freedom of thought and action and advanced physical facilities to permit maximum creative productivity. They desire broad and rapid expansion of their professional horizons with solid financial security and recognition matching personal achievements.

Our Device Development Engineer would be equally intrigued with studies in solid-state diffusion and nuclear radiation experimentation on semiconductor materials and devices. Our Circuit Development Engineer would be productively stimulated by participation in transistorization projects ranging from AM, FM and TV receivers to computer system logic.

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The cost of living is low in Nashua; schools are excellent and attractive homes are available. Downtown Boston is less than an hour away, as are the recreations of mountains, ocean, lakes.

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Through Project Engineer level. Need not be specialists, but must have creative abilities and backgrounds of VHF transmitters and receivers, communications systems in general, data processing techniques, propagation and must be capable of translating this knowledge into complex integrated systems.

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Positions Open

(Continued from page 120A)

Apply to Dr. B. G. Whitmore, Physics Dept., University of Manitoba, Winnipeg 9, Manitoba, Canada.

ENGINEER

To assist in construction of new cyclotron installation. Permanent position. Experience in building and maintaining scientific apparatus essential. Specific experience with cyclotrons desirable. Salary dependent on qualifications and experience. Apply to Dr. B. G. Whitmore, Physics Dept., University of Manitoba, Winnipeg 9, Manitoba, Canada.

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Expanding university program provides unusual opportunities for positions combining teaching and research. Rank and salary open. Teaching in all E.E. fields. Research primarily in microwave electronics. Apply: Dean E. C. Easton, Rutgers University, New Brunswick, New Jersey.

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Attractive opening for Electrical Engineer with Ph.D. to help develop graduate program, supervise graduate research and do some teaching. Encouragement given to private research and consultation. Good salary and faculty ranking will be offered to exceptional individual. Position available Sept. 1960 or sooner. Write, Dean, School of Engineering, Vanderbilt University, Nashville 5, Tenn.

ELECTRONIC ENGINEERS

Positions open to Electronic Engineers teaching lecture and laboratory courses in a rapidly growing 4 year college. A background of servomechanisms, circuit analysis, microwaves, and/or transistors is desired. Applications desired from both academic and industrial personnel. Write, Harold P. Skamser, Dean of Engineering, Calif. State Polytechnic College, Kellog-Voorhis Campus, Pomona, Calif.

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Teacher in E.E. beginning Sept. 1960. One qualified and interested in teaching fundamental undergraduate courses plus one graduate course. Ph.D. or MS. required with salary and rank dependent on qualifications. City location with many industrial contacts. Apply Dept. of E.E., Saint Louis University, St. Louis 8, Missouri.

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Research engineering position for creative person to form nucleus of a new section in the Electro-physical Labs. Will propose and evaluate new electronic component, circuit and systems concepts. Minimum experience and educational background would be 5 years with at least an MS. in E.E. or Physics. A Ph.D. degree would be preferred with strong interests in electronics, theoretical physics or mathematics. Forward resumes to Mr. R. D. Williams, Employment Mgr., P. R. Mallory & Co. Inc. 3029 E. Washington St., Indianapolis, Ind.

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Research and consulting in fields such as transportation, logistics, equipment investment, inventory control, marketing, operations research, data processing. Educational background in economics, transportation, mathematics, statistics, engineering. United Research, Inc., 808 Memorial Dr., Cambridge, Mass. Mrs. Paulsen, Personnel Director.

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The Radiation Laboratory's efforts in plasma physics include high density and low density plasmas, using plasma diagnostic techniques for the design of instrumentation to learn about the possible ionospheres of other planets, such as Venus. Analyses of possible plasma radiation from ionospheric trails, shock fronts and flame fronts are being conducted. This work is primarily in theoretical physics.

We are seeking scientists with the Ph.D. degree or with comparable significant experience in the above or closely related fields of research.

Pleasant Mid-West location with liberal fringe benefits. Salary commensurate with abilities and background.

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but we can always be good friends, can't we?
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CREDIT CARD

To the ELECTRONIC ENGINEER who neglected to MARRY THE BOSS' DAUGHTER:

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Which pretty much leaves your career up to *you*, doesn't it?

We have some advice for you; we'll not guarantee that it's impartial, but check it for logic anyway: Look for a leading electronics corporation which is essentially an engineering firm, where not only your immediate supervisors but top management will be engineers. Being engineers, they're more likely to recognize ability and to reward achievement *fairly and impartially*. It figures, we think, that where there's an atmosphere of mutual confidence, respect and understanding you'll realize your maximum potential at least a little sooner and more surely.

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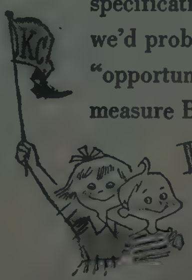

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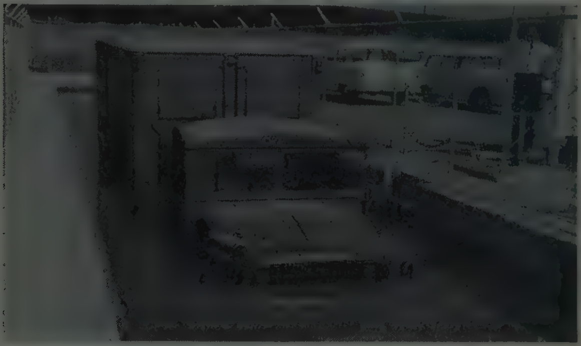
The Fording Test *is typical of the tough environmental tests imposed upon advanced electronic equipment designed and produced by Hughes Fullerton engineers.*

in the dark

TARAN (Tactical Attack Radar and Navigator) is typical of the important new electronic systems developed by Hughes—in an atmosphere famed for its engineering orientation.

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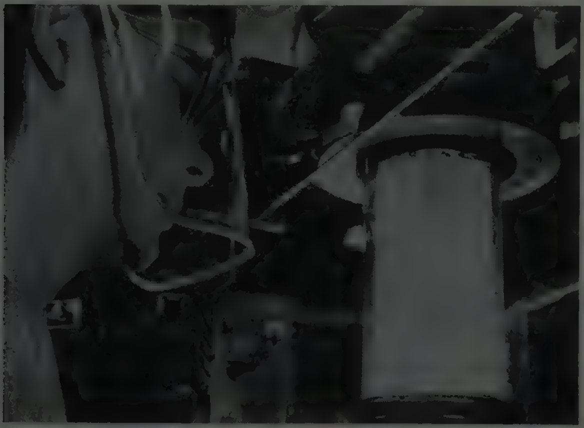
Molten Ladle *of silicon is watched during first step in the precise manufacture of Hughes semiconductors, just one of the Hughes commercial activities.*

Other Hughes activities provide similarly stimulating outlets for creative engineers. A few representative project areas include: advanced data processing systems, molecular electronics, advanced 3-D surface radar systems, space vehicles, nuclear electronics, ballistic missiles, infrared devices—and a great many others. The commercial activities of Hughes have many interesting assignments open for imaginative engineers to perform research, development, manufacturing of semiconductors, electron tubes, and microwave tubes.

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Hughes General Offices, Bldg. 6-E4, Culver City, Calif.

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You will serve as liaison between Raytheon and licensees, assisting in the design of microwave cooking units for these licensees' commercial ovens. This will involve working with the licensees' design engineers to solve such problems as heat, sanitary considerations, microwave distribution, wave modulation, insulation and cost reduction. You will also help translate licensees' suggestions into product improvements.

You should be a senior-level electronics engineer with a BSEE or equivalent and should have 5 years' applicable experience. Background in design of test equipment, high-power modulators, or commercial microwave equipment is especially useful, as is knowledge of magnetrons, underwriters' requirements and FCC regulations. Basically, we're looking for a top design and sales engineer (travel will be limited, however) so salary is commensurate—that is, excellent.

Both positions are in the suburban Boston area and offer attractive cultural, educational and recreational opportunities. Relocation assistance is available. Please phone collect (BIgelow 4-7500) or forward your detailed resume to Mr. Paul R. Alexander, Manager of Professional Personnel, Commercial Apparatus & Systems Division, Raytheon Company, Watertown, Massachusetts.

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By Armed Forces Veterans

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The IRE publishes free of charge notices of positions wanted by IRE members who are now in the Service or have received an honorable discharge. Such notices should not have more than five lines. They may be inserted only after a lapse of one month or more following a previous insertion and the maximum number of insertions is three per year. The IRE necessarily reserves the right to decline any announcement without assignment of reason.

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BSEE., extensive antenna and R.F. experience, design and project supervision. Broad background includes sales and engineering management, contract negotiations. Desires position as General Manager of antenna company, or participant in ownership management of antenna consulting firm or small growing company. Box 2054 W.

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E.E. degree. 10 years solid experience in electromechanical components—development, evaluation, application and standardization. Desires a challenging, responsible position in a supervisory capacity, New York, Long Island area preferred. Box 2055 W.

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(Continued on page 128A)

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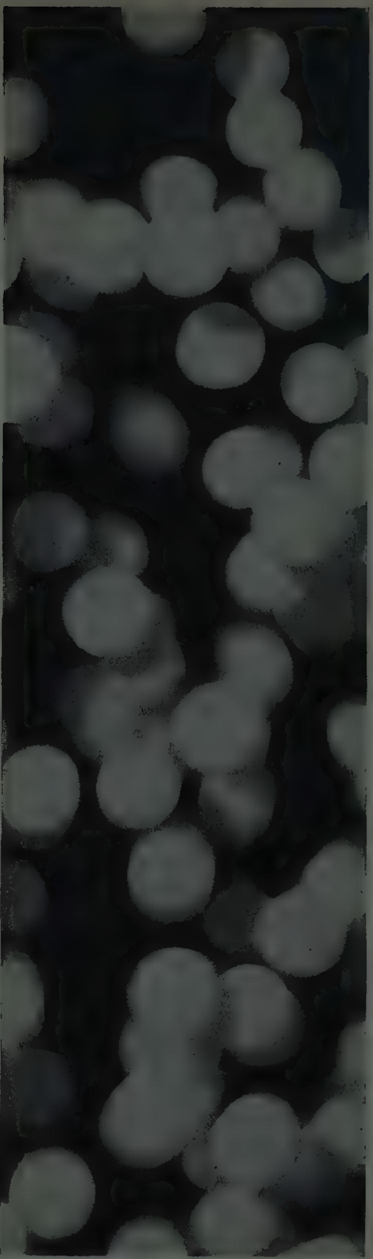

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Positions Wanted

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(Continued from page 126A)

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BSEE. 1957. 1/Lt. Sig. C. 1½ years as Project Officer at WSMR, New Mexico. Experience in the field of guidance and countermeasures of missiles. Interested in obtaining a LL.B. Desires position which requires an engineering and legal combination with management opportunities. Box 2072 W.

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ENGINEERING MANAGER

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(Continued on page 130A)

KNOWLEDGE

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(Continued from page 128A)

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ENGINEERING MANAGER

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ELECTRICAL ENGINEER

5 yr. E.E. plus good Liberal Arts background. Polyglot; age 31; family; lifelong U.S. citizen. Diverse experience ranging from power during military, 3½ years design and manufacture elec-

(Continued on page 133A)

ELECTRONICS ENGINEERS
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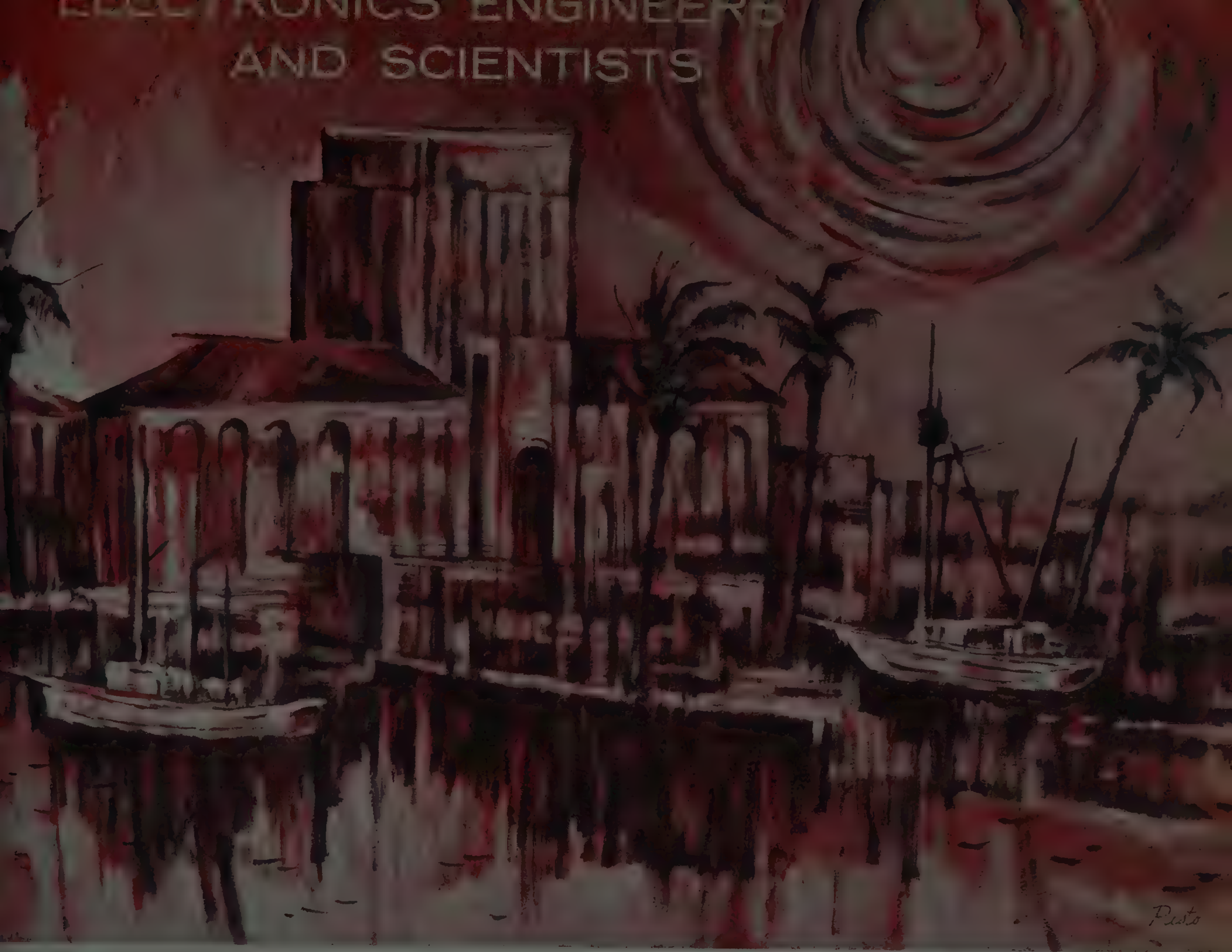

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THE ARTIST — Sammy Pasto, Convair-San Diego artist whose painting appears on the preceding page, is a native of Boston. He studied at the New England School of Art and moved to California following World War II. Presently working as an illustrator at Convair-San Diego, Pasto has received broad recognition for his work in fine arts. He has had one-man shows in New York, Philadelphia and Los Angeles and his work has been accepted in most juried West Coast art shows. For a change of pace, Pasto plays clarinet in a prominent Dixieland jazz band in San Diego.

PROFESSIONAL EMPLOYMENT INQUIRY — This information will enable the professional staff at Convair/San Diego to make a preliminary evaluation of your background. A personal interview can be arranged in your city by appointment. Every completed inquiry will be acknowledged.

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Other________________
EXPERIENCE: Primary Specialty________________
Secondary Specialty________________
YEARS OF EXPERIENCE IN PRIMARY SPECIALTY:________
Additional Comments, if any, concerning your job interests________

Positions Wanted

By Armed Forces Veterans

(Continued from page 130A)

tric electronic instrumentation and control, 2 years solid state R and D, 1 year specifications and estimating for contracts and projects, 4½ years supervisory, presently project level problem solver and trouble shooter for large company on communication, navigation, ultrasonic and control. Box 2084 W.

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(Continued from page 112A)

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Schiowitz, H. H., Canoga Park, Calif.
Schneider, P. C., North Caldwell, N. J.
Schulkind, D., Massapequa, L. I., N. Y.
Schumacher, T. J., Cincinnati, Ohio
Schwartz, S. S., Asbury Park, N. J.
Schwarz, E. G., North Syracuse, N. Y.
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(Continued on page 134A)

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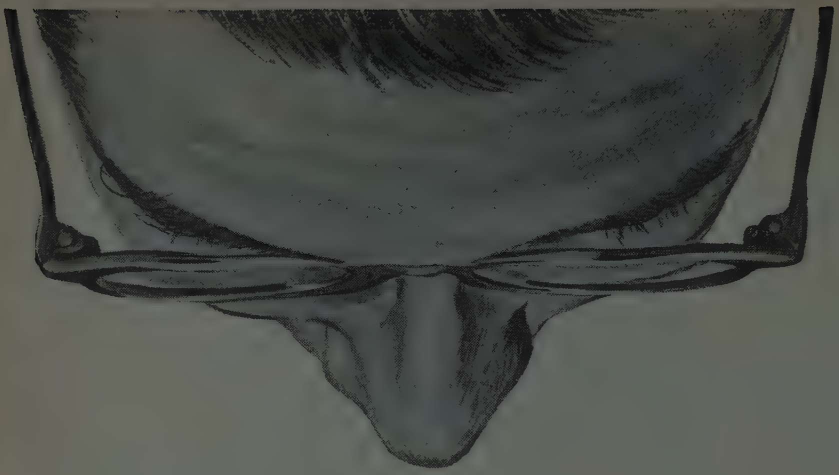

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(Continued on page 136A)

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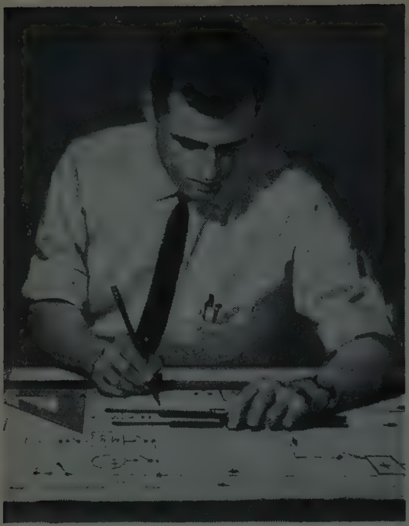

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(Continued on page 138A)

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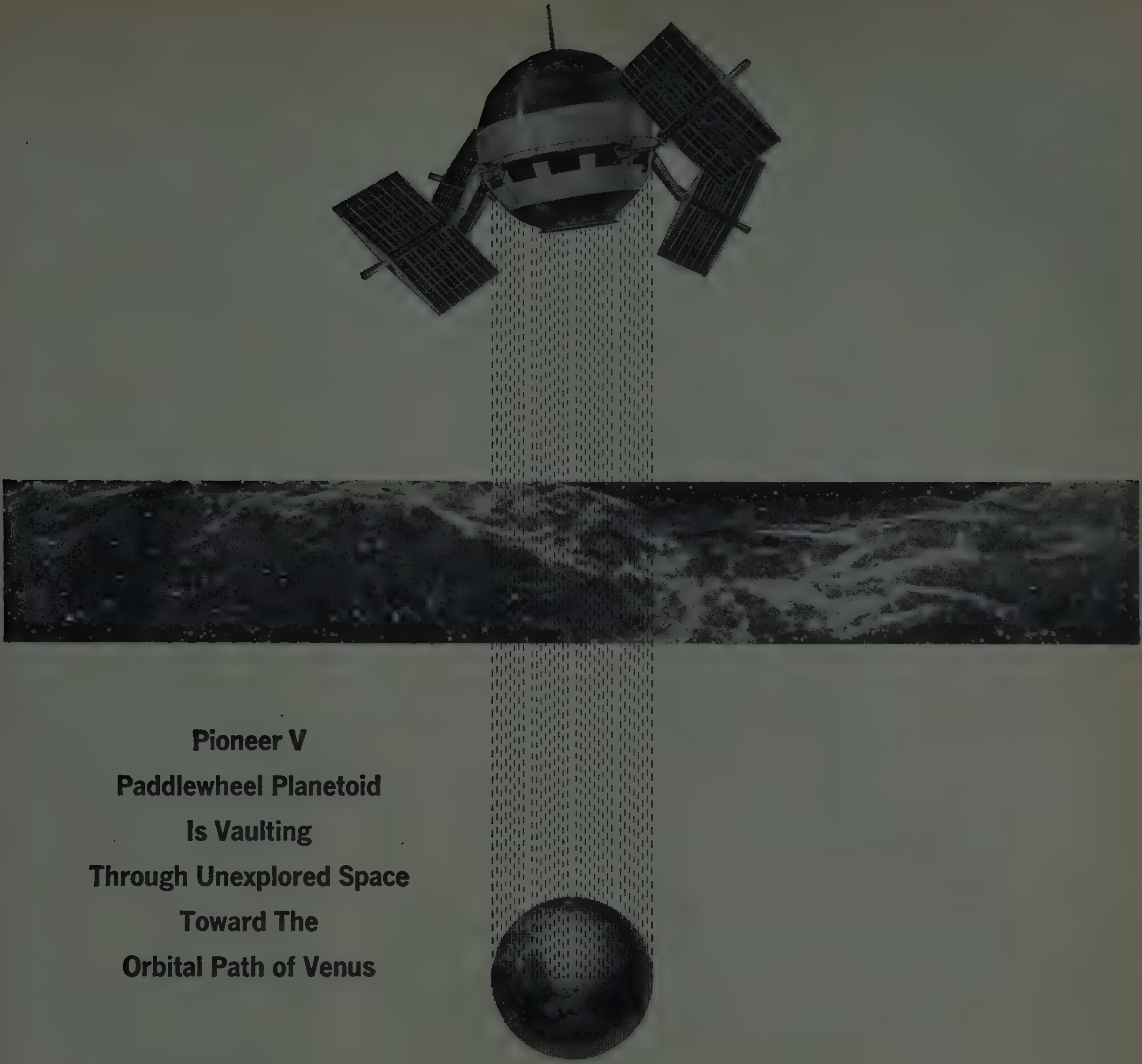

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(Continued on page 146A)

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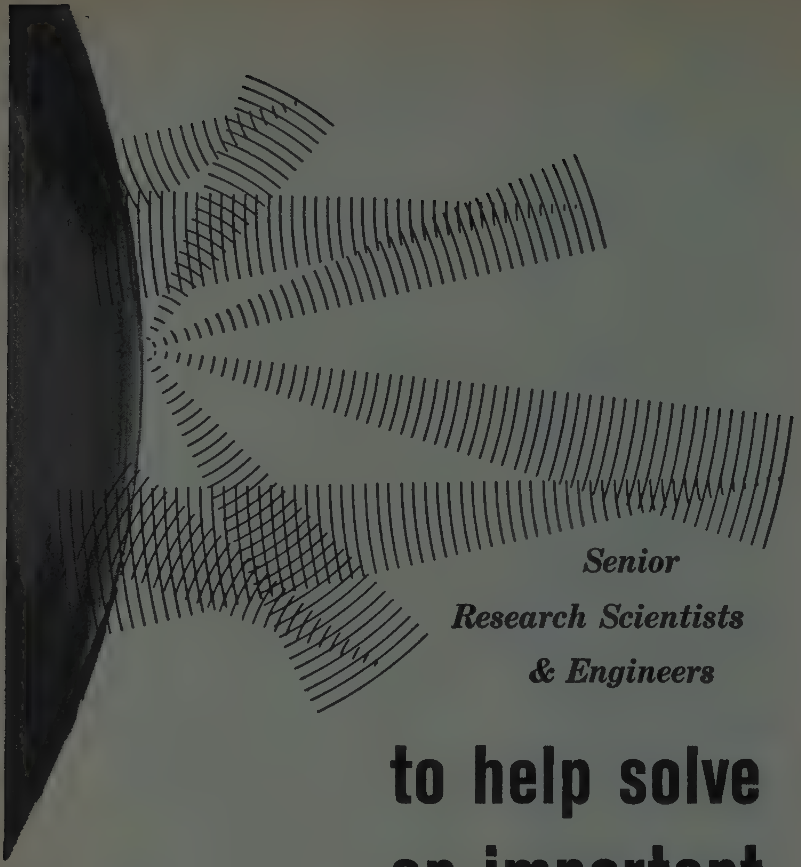

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(Continued from page 144A)

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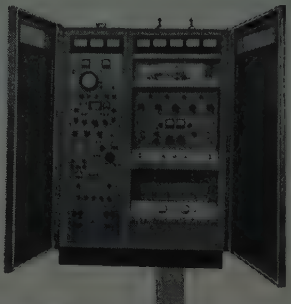

GPR-90
Bulletin
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Our Tactical Systems Laboratory applies advanced techniques to the design and development of airborne and ground-based digital data processing systems. If you have at least 2 years of design, system integration, testing or production experience in digital systems, your talents may find application in the solution of our technical problems. Write to Mr. S. L. Hirsch.

LITTON INDUSTRIES Electronic Equipments Division
Beverly Hills, California

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 42A)

Umbilical Connector

Automation Electronics, Inc., an operating division of **Arnoux Corp.**, 11924 West Washington Blvd., Los Angeles 66, Calif., released data on its umbilical connectors which have been under development for some months. Already in use on an intermediate range missile, the unit is small, lightweight, and of integrated construction. The umbilical also features a mating and unmating mechanism so that the whole operation may be accomplished without the use of any tools.

This particular design provides for 300 individual contacts, each rated at 5 amperes and all are mounted in a space 2¾ inches square. The whole unit is sealed in a magnesium aluminum casing which matches a cutout in the missile skin. There are two fluid lines, each capable of carrying three gallons per minute with a maximum pressure drop of 6 psi. Disengagement is accomplished in two ways: normal motion of the missile disengages these contacts; and the plug may also be removed manually without missile motion.

The contacts are Multifilar and consist of heat-treated beryllium copper spring wires, which are silver-flashed and gold-plated. This design provides for high resistance to corrosion. Another advantage of this kind of contact is that a great number of disconnects are possible with very little degradation of the electrical contacts.

Further information may be obtained by writing to Arnoux.

Miniature Potentiometers

A new series of ⅝" diameter miniature compact composition variable resistors is offered by **Chicago Telephone Supply Corp.**, Elkhart, Ind. Series 200 can be furnished with or without attached switch in standard bushing mounted construction or in the more economical ear mounting. Resistance range is 250 ohms through 2.5 megohms linear taper, wattage rating ¼ watt through 100,000 ohms and 2/10 watt over 100,000 ohms at 55°C derated to no load at 85°C, voltage rating 750 vac bushing to terminals for 1 minute high pot test and 500 vdc operating maximum, 350 vdc across end terminals and 280° rotation without switch, 315° with switch. De-

(Continued on page 150A)

General Motors pledges
AC QUESTMANSHIP
2

AC Seeks and Solves the Significant—Inspired by GM's pledge to contribute heavily to our national defense, AC, an acknowledged leader in the new technology, plans to reach far beyond such accomplishments as AChiever inertial guidance systems. / This is AC QUESTMANSHIP. It's an exciting scientific quest for new ideas, components and systems . . . to promote AC's challenging projects in guidance, navigation, control and detection. / Mr. Jack Briner, AC Director of Field Service, believes his department's Career Development Program "offers young engineers world-wide opportunities in the practice of Questmanship." They learn a product from its technological theory through its operational deployment. Following this training, "they utilize their own ingenuity to support AC products in the field, with more effective technical liaison through training, publications, maintenance engineering, and logistics." / You may qualify for this special training, if you have a B.S. in the electronics, scientific, electrical or mechanical fields. Special opportunities also exist at AC for men with M.S. and Ph.D. degrees. If you are a "seeker and solver," write the Director of Scientific and Professional Employment, Mr. Robert Allen, Oak Creek Plant, Box 746, South Milwaukee, Wisconsin.

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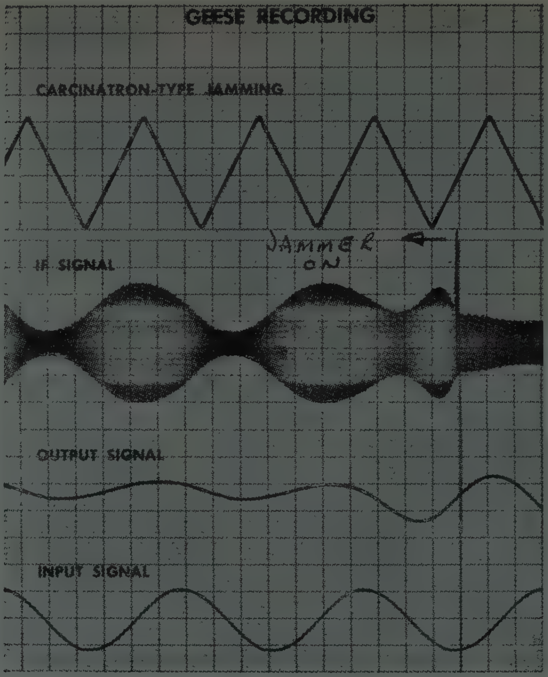

EVOLVING LARGE-SCALE SYSTEMS CONCEPTS

AND DEVELOPING THE TOOLS THAT SPEED THEIR DESIGN CYCLE

Defense Systems Department is directing its technical capabilities toward the development of large-scale electronic systems. Inherent within this work program is the recognition, definition and solution of problems in every aspect of the systems technology.

To accomplish this ambitious task, a growing number of studies are being directed toward the development of unique tools that will aid in the design of superior systems in less time, at lower cost.

A recent contribution by Defense Systems Department in this technological area is GEESE (General Electric's Electronic System Evaluator). Utilizing advance computer techniques, it enables systems engineers to accurately predict, optimize and synthesize system performance prior to design.

GEESE is indicative of the scope of Defense Systems Department's involvement in the systems technology. Many programs offer systems-oriented engineers and scientists an opportunity to participate in new areas of long-term importance.

Senior members of our technical staff would welcome the occasion to discuss personally and in detail the career positions available with this growing organization. Address your inquiries in professional confidence to Mr. E. A. Smith, Box 4-F.

DSD DEFENSE SYSTEMS DEPARTMENT
A Department of the Defense Electronics Division

GENERAL ELECTRIC

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NEWS New Products

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(Continued from page 148A)

signed for commercial applications where space is at a premium. For applications requiring extremely compact control and SPST switch combinations, unique Type QS-200 combined switch and control in a single miniature molded housing. Base prices in 3000-9999 quantities for the various Series 200 constructions start at $160.42 M without switch and $316.93 M with switch. Delivery 4 to 5 weeks after sample approval.

Variable Transformers

Individual transformers operating from a single-phase, 240-volt line are now available from **Ohmite Manufacturing Co.**, 3696 Howard St., Skokie, Ill. Model VT8H is rated 3 amperes and at maximum setting will deliver line voltage (240v) or overvoltage (280v). Model VT8HN is the "no-overvoltage" type which will deliver just line voltage at maximum setting, but offers a current bonus in its output rating of 4 amperes.

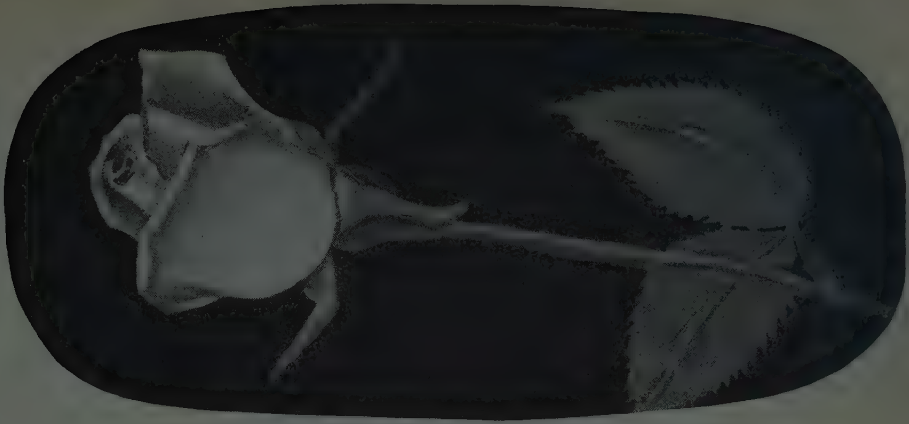
Both models are constructed on the same frame as Ohmite's popular VT8 series (ratings for 120-volt input types; 7.5 amperes with overvoltage; 10.0 amperes without overvoltage). The "high voltage" units afford the convenience of 240-volt, single-phase operation using a single unit instead of a ganged assembly. The 240-volt, single-phase operation can also be achieved with ganged assemblies of 120-volt input transformers where higher power output is desired.

The transformers have a tap which permits operation at 120 volts input with 280 volts output for the VT8H and 240 volts output for the VT8HN (some derating is required for output beyond 120 volts). These high voltage transformers can also be obtained in ganged assemblies to achieve 240-volt, 3-phase operation or 480-volt single phase or 3-phase operation. Request Bulletin 151.

Sweep Generator

A new Sweep Generator is now available from **Telonic Industries, Inc.**, Beech Grove, Ind., with twice the sweep width of

(Continued on page 152A)



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There are many worthwhile advantages awaiting the Professional Engineer who chooses to advance his career with us at Bendix York.

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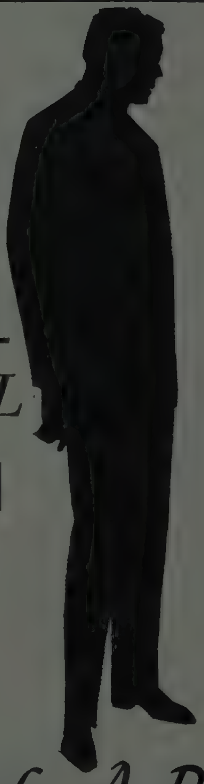

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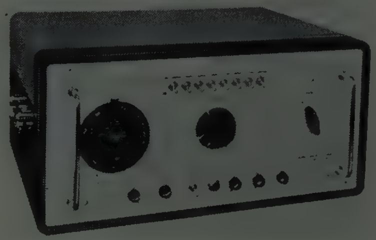
(Continued from page 150A)

earlier models having the same range and power output. The new Model HD-1A, has a center-frequency range of 1 to 900 mc, and a sweep width that is adjustable up to 200 mc in the lower portion of the range.



The wide range is achieved by using two UHF swept oscillators, each with a range of approximately 400 to 900 mc. These are combined in a heterodyne circuit to produce a third signal that is tunable over a 1-to-400 mc range. The output of a single oscillator is used for the instrument's upper range, providing a continuous spectrum of center frequencies from 1 to 900 mc.

Sweep widths may be varied from 200 mc in the heterodyne portion of the center-frequency range. Over 400 mc, the sweep width is variable from 0.06 to 10% of the center frequency. Attenuators, detectors and other circuit elements are carefully designed for broadband performance so that the output is level across the entire swept frequency, even when set at the widest limit. The two oscillators are leveled by an automatic-gain-control circuit that continuously samples the swept output. Specifications for the HD-1A call for a flatness of less than ±5% at maximum sweep width.

The RF output of the instrument is a maximum of 0.75 volts, peak-to-peak, across the heterodyne range and 2.0 volts, peak-to-peak, in the high range. These voltages are adjustable by a turret attenuator mounted on the front panel of the instrument. Settings are provided for 0, 10, 20, 30, 40, and 50 db attenuations, plus a 0–10 db vernier.

By automatically blanking the return sweep, the Model HD-1A instrument provides a base line on the scope presentation. The horizontal sweep signal is 20 volts, peak-to-peak, with a repetition rate equal to line frequency, 50 to 60 cps.

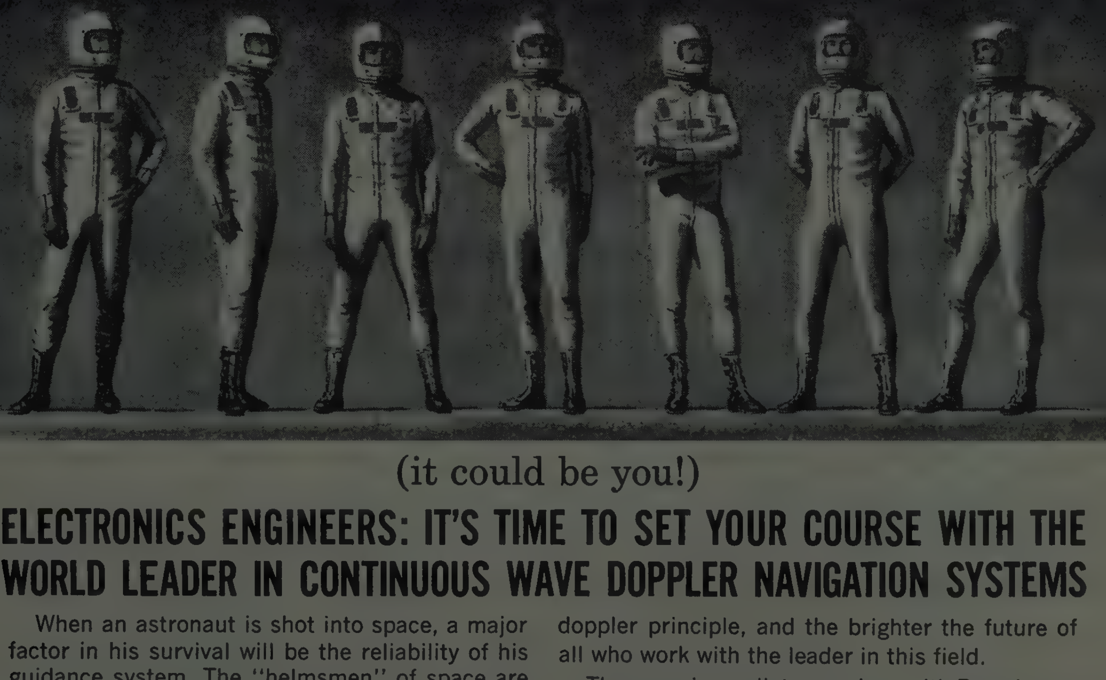
Toggle switches on the face of the instrument introduce fixed "birdy" markers into the oscilloscope trace. Up to eight plug-in marker units may be used, each set at the factory to a particular frequency. In addition, harmonic plug-in markers are available that add birdies at 1 mc, 10 mc or other intervals.

High Temperature Precision Capacitors

A new line of precision film capacitors is now available in the −55°C to 125°C

(Continued on page 155A)

ASTRONAUTS IN SEARCH OF A HELMSMAN



(it could be you!)

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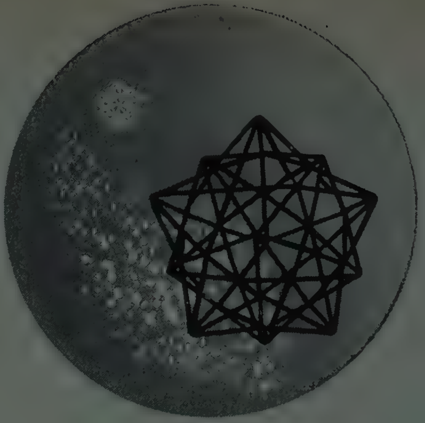

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Technical Employment Manager,
Department 14D

REPUBLIC AVIATION

Farmingdale
Long Island, New York

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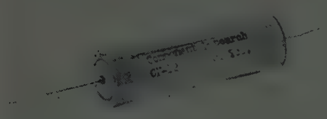
(Continued from page 152A)

temperature range from **Component Research Co., Inc.,** 3019 So. Orange Drive, Los Angeles 16, Calif.



The capacitance change is less than 1% for the entire range. From 25 to 125°C the capacitance change is less than ±0.3% and zero ±30 ppm /°C temperature coefficient, which remains stable with repeated temperature cycling.

Insulation resistance is greater than 10^{13} ohms. Dielectric absorption is less than 0.0003 when measured with a charging voltage of 44 volts for 30 seconds.

The dissipation factor remains less than 0.0003 over the entire temperature range and a wide range of frequencies.

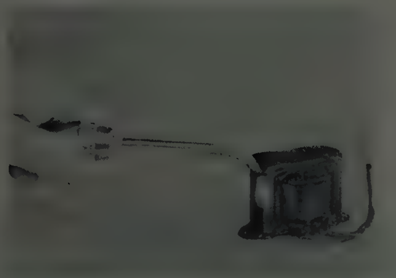
These CR PROCESS (pat. pend.) capacitors have a standard tolerance of 2% and are hermetically sealed in tubular and bathtub enclosures. Other tolerances ranging to 0.1% are available at all standard working voltages.

Special capacitors with less than ±0. 3% total capacitance change from −80°C to 125°C can be supplied for unusually critical circuit applications.

These precision capacitors can be used in computers, timing circuits, servo systems, pulse filter networks and a variety of other applications.

Fire Retardant Casting Resin

Emerson & Cuming, Inc., Canton, Mass., announces a new fire retardant epoxy casting resin which may be cured at room temperature. The material has excellent adhesion to metals, plastics and ceramics.



Due to its low thermal coefficient of expansion and high heat stability, large embedments can be made. It is stable over a temperature range of −100 to +350° F. Stycast 1231 can be color coded to specification but normally is supplied in black.

(Continued on page 156A)

MAN OUT...

A fascinating project at Martin-Denver and one which offers to the truly creative engineer or scientist a personal esteem and professional recognition unequalled in today's opportunities. Please do consider being a part of this or other creative involvements at Martin-Denver and inquire of N. M. Pagan, Director of Technical and Scientific Staffing, (Dept. DD 9), The Martin Company, P. O. Box 179, Denver 1, Colorado.

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development engineers

The Brown Instrument Division of Honeywell has several openings for development engineers. Work involves responsibility for projects from drawing of the original requirement specifications, through the development and design phases, until finished drawings and specifications are released to the manufacturing department.

JR. & SR. ELECTRONICS ENGINEERS

Engineering grads with at least 1 year in electronics development work. Present or contemplated projects include development of high gain, low level d-c servo amplifiers; computing elements for industrial process control loops utilizing transistor and magnetic amplifier circuitry; sophisticated controllers for process control utilizing diode modulator techniques with a high audio frequency carrier; and high gain, wide band d-c amplifier development.

JR. & SR. ELECTRICAL ENGINEERS

BSEE's with at least 1 year of experience in the development of electrical measuring instruments such as potentiometer recorders with microvolt sensitivity, galvanometer indicating controllers, current actuated recorders, etc. Also involves development of many types of electrical transmitters and components such as strain gauges, galvanometers, slidewires, torque motors, and electro-mechanical transducers. Frequently the work on these instruments involves telemetering components, integrators, controllers and alarm devices.

ELECTRICAL PRIMARY MEASURING ELEMENTS ENGINEERS

BSEE's or Engineering Physics grads, with at least 3 years' experience, and interested in electrical output signal. Develop new products, investigate and correct unusual production and field troubles and keep in close contact with competitive developments.

If you are qualified and a U.S. citizen and interested in any of these openings, send information on your background and experience to C. R. Brandt, Employment Manager, Dept. 544,

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Brown Instruments Division

Wayne and Windrim Avenues, Philadelphia 4, Pa.

To explore professional opportunities in other Honeywell operations coast to coast, send your application in confidence to B. I. Eckstrom, Honeywell, Minneapolis 8, Minnesota.

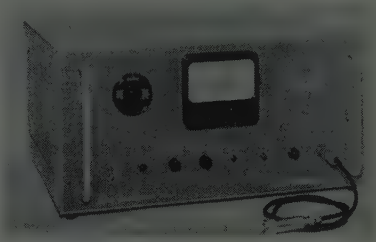
NEWS New Products

(Continued from page 155A)

Thin sections of cured Stycast 1231 repeatedly introduced to a flame will, on each occasion, immediately extinguish.

Noise Source

The Mega-Node 3000, a noise-diode type calibrated noise source, providing low VSWR across a frequency range of 1 to 3000 mc, has been introduced by the **Kay Electric Co.**, Dept. PI, Maple Ave., Pine Brook, N. J., manufacturers of precision electronic test and measuring instruments.



Accurate noise-figure measurements require diode current to be known and generator impedance to be constant with frequency, since available noise power is a function of both. In the Mega-Node 3000, the generator impedance is held constant across the range by an impedance matching device that is simple in principle. Thus, the error introduced by the generator

(Continued on page 158A)

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- Telemetry
- I.F. & Video Circuitry
- Environmental Testing

STAVID's facilities in Plainfield, New Jersey, at the foothills of the Watchung Mountains, are near excellent schools, modern shopping facilities and ample housing accommodations. With New York City just 45 minutes away, and the New Jersey shore within one hour's drive, the Plainfield area provides an ideal environment for work, recreation and comfortable suburban living.

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Experienced in actual sales, installation and servicing of engineering and business data processing systems in major industrial concerns.

Systems Analysts

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(Continued from page 156A)

VSWR (impedance variation) is minimized.

Rapid, accurate measurement of noise-figure can be accomplished without calculation of the effective bandwidth. Variable noise-power range and low VSWR permits accurate noise-figure measurements. Calibration charts and slide rule are not required to determine noise-figure.

A coaxial-type, tungsten-filament noise diode is employed as a temperature-limited noise-generator. Plate voltage is regulated and diode plate current, which is proportional to noise power output, is adjustable by a front panel control. Current is indicated on the panel meter, calibrated in noise-figure. To prolong the life of the noise diode, an automatic cutoff is incorporated to limit the time of operation at high currents.

The panel meter reads noise-figure directly up to 20 db for a 3 db Y-factor. Maximum VSWR is 1.15 from 1 to 250 mc; 1.25 from 250 to 2000 mc; 1.4 from 2000 to 3000 mc. Noise-figure accuracy is ±0.25 db from 1 to 250 mc. At the higher frequencies, the noise-figure error (db) is a function of the generator VSWR, but does not exceed SWR (db)/2.

Maximum VSWR indicated occurs at frequencies which represent less than 20% of the total frequency range. The VSWR over most of this range is much lower than the maximums specified.

Specifications: Frequency Range, 1 mc to 3000 mc; Output Impedance is Nominal 50 ohms (type-N connector); Noise Figure Range, is 0 to 20 db; Meter Calibration, Logarithmic in db noise-figure and linear in dc ma; Noise Diode, Bendix Type TT-1. Power Supply, Input approximately 200 watts, 117-v (±10%), 60-cps (Available for 50-cps operation on special order). B+ electronically regulated. Dimensions, $9\frac{3}{4}'' \times 19\frac{1}{2}'' \times 15\frac{1}{2}''$; Weight: 44 lbs.

Transistorized Frequency Changers

ERA Pacific, Inc., 1760 Stanford St., Santa Monica, Calif., announces new additions to its line of transistorized frequency changers. These units convert one input frequency to a second frequency at power levels and eliminate the disadvantages inherent in vacuum tube or mechanical equivalents. The new units are designed for powering all types of AC equipment, gyro and servomechanisms, magnetic amplifiers and are intended for laboratory and industrial applications.

(Continued on page 163A)

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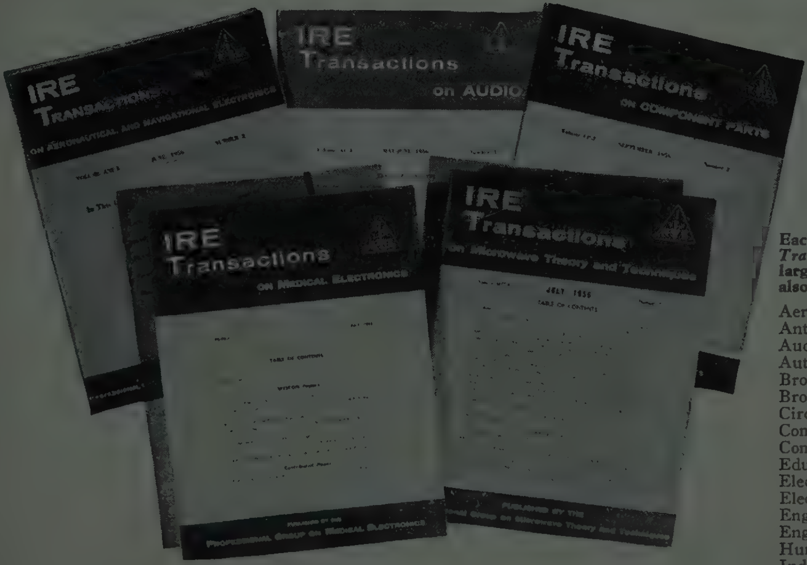
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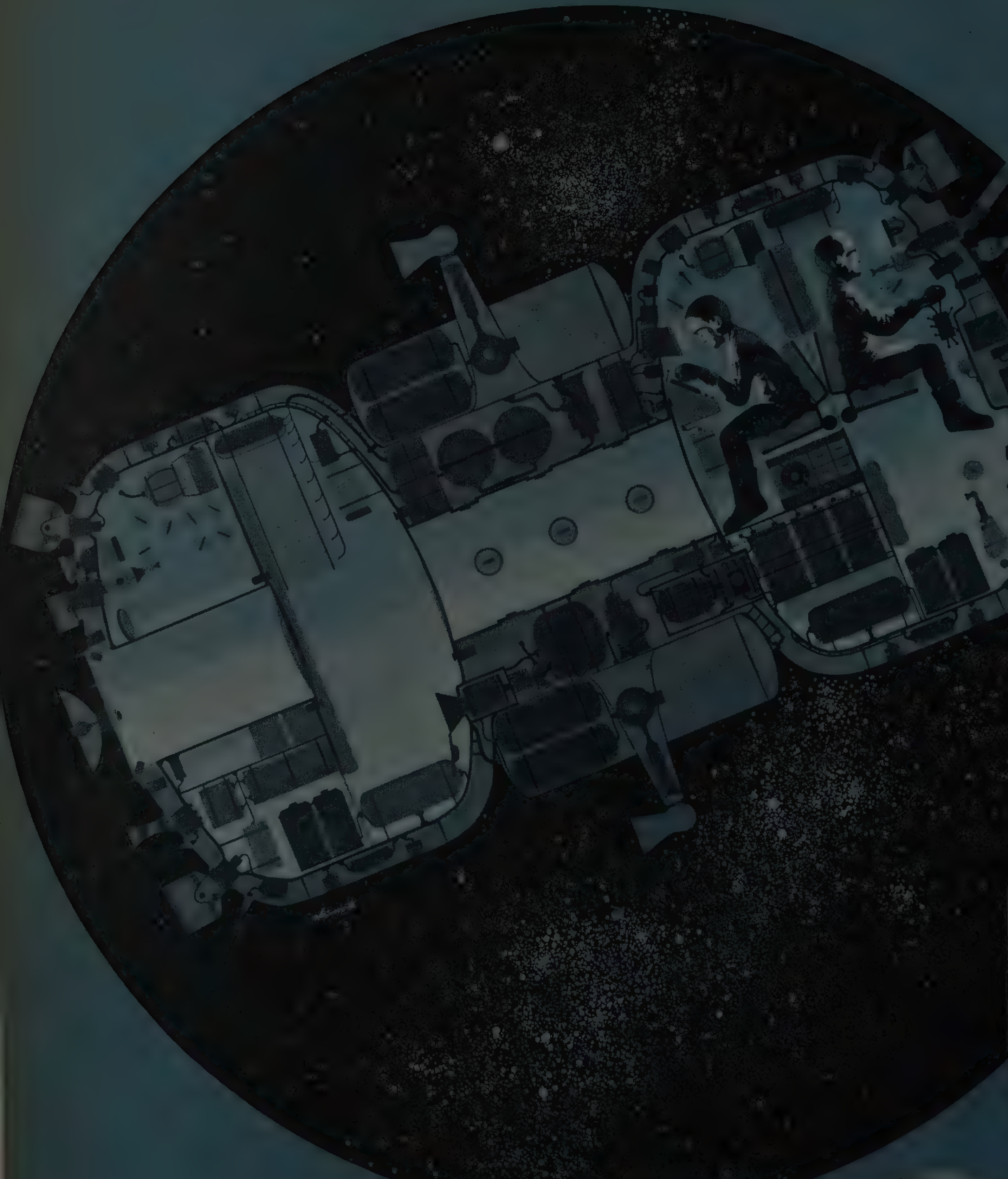
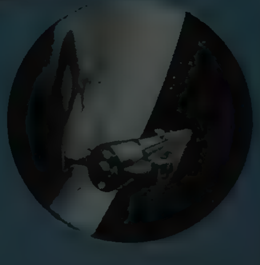
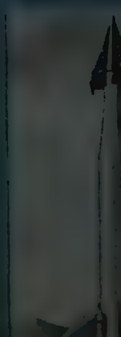






THE

STROTUG

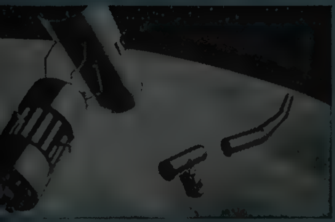
Tugboat for Space: Spaceborne scientific laboratories and platforms for further exploration into space are an accepted concept based on established engineering techniques. Components would be fired as individual units into space, on precalculated orbits, and there assembled. To solve the major problems of how men are to live and work in space during the assembly process, Lockheed has prepared a detailed engineering design of an astrotug—a manned vehicle housing a crew of two or three. Missile-launched, the astrotug will be capable of supporting its crew for a number of days in an environment of suitable atmosphere, artificial gravity, and with provisions for exercise, relaxation, bathing facilities, medical care, illumination and adequate food and water.

The Lockheed astrotug is a completely independent working vehicle. Personnel need not leave it in space suits in order to work on the project of assembling the space station components. As shown in the diagram, the tug consists of two double-walled pressure vessels approximately 20 feet long overall and 9 feet in inside diameter. Swivelling rocket nozzles are arranged for maneuvering. On the forward end, extending out are four mechanical manipulator arms with interchangeable "hands" for such specialized functions as gripping, welding, hammering, cutting, running screws, etc. "Hands" can be changed by remote control from inside. Viewing ports provide uninterrupted observation. Radar antennas, searchlights, and other equipment necessary to the tug's work are mounted externally. Main controls and instruments including radar, radio, infrared, computers and navigation consoles are duplicated in each of the two major compartments as a safety measure.

Men working in single units afloat in space suits would have little applicable force and could work for very limited periods of time. With the Lockheed astrotug, personnel could carry on the work in relative safety and comfort with maximum efficiency. A special reentry vehicle, separate from the astrotug, has been conceived for ferrying to and from earth. Tugs themselves would remain floating in orbit indefinitely, being reprovisioned and refurbished as fresh crews arrive in relief.

Space vehicle development is typical of Lockheed Missiles and Space Division's broad diversification. The Division possesses complete capability in more than 40 areas of science and technology—from concept to operation. Its programs provide a fascinating challenge to creative engineers and scientists. They include: celestial mechanics; computer research and development; electromagnetic wave propagation and radiation; electronics; the flight sciences; human engineering; magnetohydrodynamics; man in space; materials and processes; applied mathematics; oceanography; operations research and analysis; ionic, nuclear and plasma propulsion and exotic fuels; sonics; space communications; space medicine; space navigation; and space physics.

Engineers and Scientists: Such programs reach far into the future and deal with unknown and stimulating environments. It is a rewarding future with a company that has an outstanding record of progress and achievement. If you are experienced in any of the above areas, or in related work, we invite your inquiry. Please write: Research and Development Staff. Dept. D-33, 962 W. El Camino Real, Sunnyvale, California. U.S. citizenship or existing Department of Defense clearance required.



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To design, execute and interpret experiments in vacuum tubes. Ability to design and build electronic circuits for measuring very small currents necessary. Broad experience in vacuum tube technology and preference for experimental work highly desirable.

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To perform theoretical analysis and associated experiments in electron and ion dynamics. Broad experience in field of vacuum technology and familiarity with metallurgy and materials in vacuum tube technology highly desirable. Capable of assuming technical direction and responsibility for challenging projects.

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These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 158A)

Three new models are available—the Model FC 6415 which converts an input frequency of 60 to 400 cps at a 150va power rating, the Model FC 6410 which converts a 60-cycle input to a 400-cycle output at 100 va, and the Model FC 645 at 50va. These models provide a regulated output frequency compensated against both input and load variations.

The new models achieve frequency conversion by means of special circuitry which converts the input frequency to dc by means of silicon rectifiers. This dc, in turn, is converted to the desired output frequency by means of stabilized transistor keyers. This design approach permits desirable achievements, such as instant starting, small size, light weight, zero maintenance, negligible radio interference and high shock vibration resistance characteristics.

Contact-Making Wattmeter

A new kind of ac-dc contact-making wattmeter, capable of both indicating and controlling power in the lower milliwatt range, is announced by **Assembly Products Inc.,** Chesterland, Ohio.

(Continued on page 165A)

gp

Now that you're back...

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R & D Operations Liaison. Must have demonstrated technical ability to direct the incorporation of modern ideas into field equipment: to actively participate in the equipment transition from prototype to operational status. These are senior appointments as liaison between our R & D Equipment Development Laboratory and Field Operations. Degree in E.E. and the ability to communicate effectively are required.

junior electronics engineers

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The Electro-Physics Laboratories are located in the suburban Washington, D.C. area, where post-graduate study is available in several nearby universities. Housing is plentiful in attractive, well-established neighborhoods. Our relocation allowance is liberal. Inquiries are held in complete confidence. A prompt reply will follow the receipt of your resume by:

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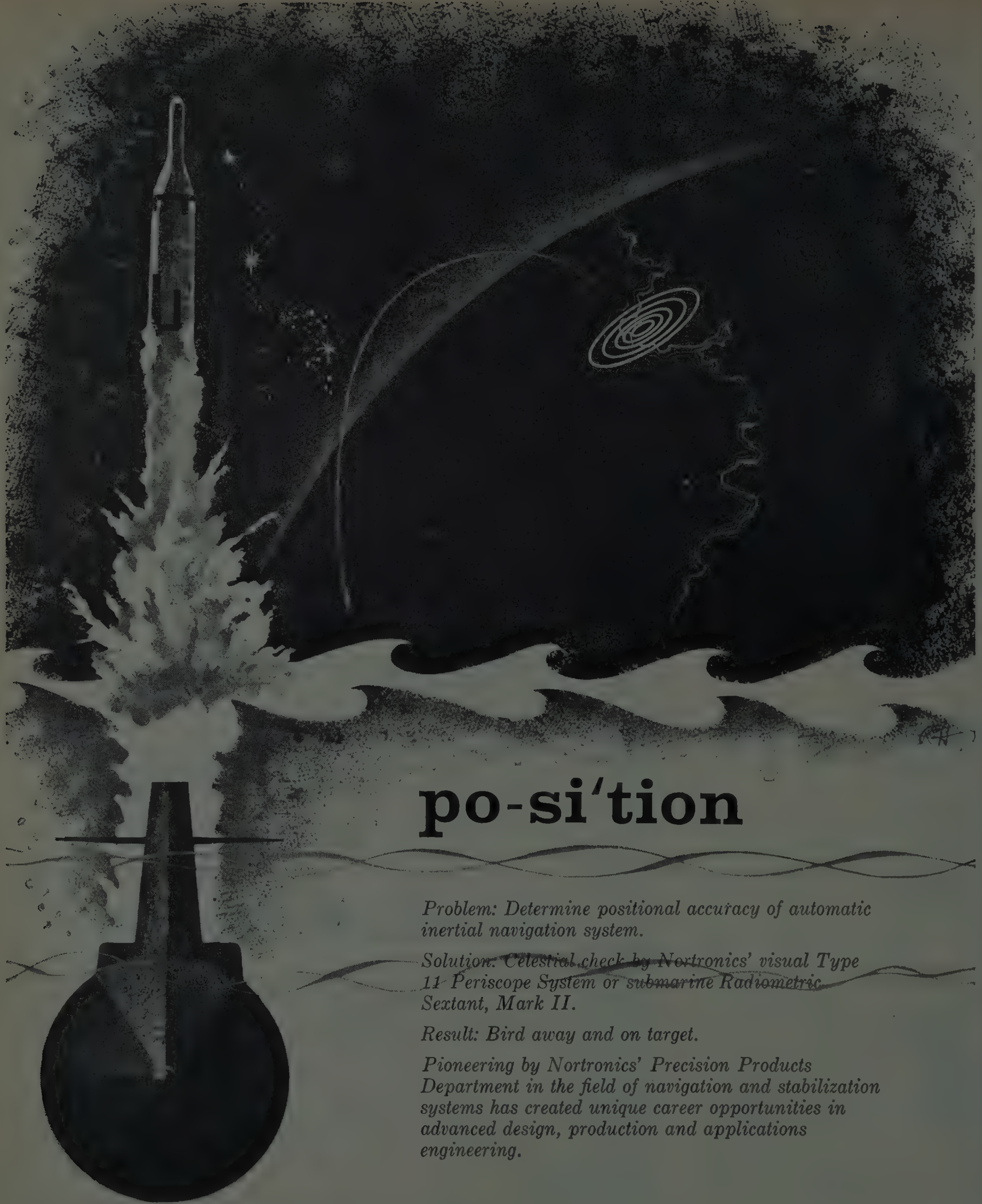

po-si'tion

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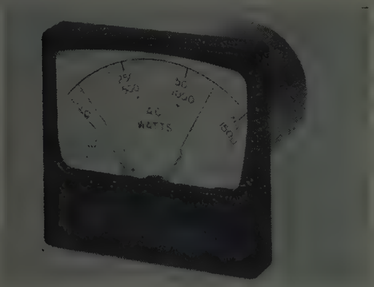
These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 163A)

Although primarily intended for measuring power, the unit also is available in modified form as a true RMS voltmeter or ammeter, or as a varimeter. The instrument can monitor the product of as many as four different variables.

Main elements of the instrument are a locking contact meter-relay and a magnetic circuit built around a Hall effect solid state device. They come in a panel-mounting package with a 4½-inch meter and a barrel less than 5 inches long.



Greatest standard full scale sensitivity at present is 500 milliwatts, either ac or dc, but with modifications dc ranges as sensitive as 100 milliwatts are available. The instrument can be built to operate on any ac frequency up to 1000 cps. Either ac or dc must be specified—they cannot be measured interchangeably by a single instrument.

Power measurement is simplified by eliminating the standard dynamometer movement of wattmeters. This is possible because the Hall device puts out a voltage that is proportional to the power in a load circuit. The output voltage is fed to a standard D'Arsonval movement that is directly calibrated for power measurement.

As with standard API locking coil meter-relays, control action begins whenever the signal pointer reaches a limit set by either a high or low adjustable pointer. Circuits are then closed that actuate load relays and other control components.

The new wattmeter is available with any of the standard API circuits used with locking contact meter-relays to provide a variety of control patterns. One of the more obvious applications is controlling motor load, since power is a more reliable measurement factor than the current alone that is frequently used.

Where remote indication or control adjustment is desired, the meter-relay and the associated circuitry are furnished in two packages for separate mounting.

Silicon Mesa Power Transistor

Pacific Semiconductors, Inc., 10451 W. Jefferson Blvd., Culver City, Calif., announces very high frequency silicon mesa power transistors capable of delivering one watt power output at 70 mc with a 28 volt collector voltage.

The two new types, designated as 2N1505 and 2N1506, are characterized by 3 watt collector dissipation, 40 volt collector to emitter rating, and low collector capacitance. The units are suited for VHF power application in communications and telemetry equipment where severe environmental conditions are encountered.

Type 2N1505 operates as an oscillator at 70 mc with a power output in excess of 1 watt at an efficiency of 45%.

Type 2N1506 has a typical power gain of 12 db at 70 mc with a useful power output of 1.0 watts. At 200 mc the 2N1506 has a power output of 300 milliwatts.

Delivery is currently being made on both new types. The 2N1505 is priced at $39.50 up to 99 pieces at $29.60 up to 1000 pieces. The 2N1506 is $49.50 and $37.00 respectively.

The new types supplement the present PSI line consisting of high frequency, high power, high voltage types 2N1335–1341, the multipurpose high speed switches 2N1409–1410 and medium power high speed types 2N696/697.

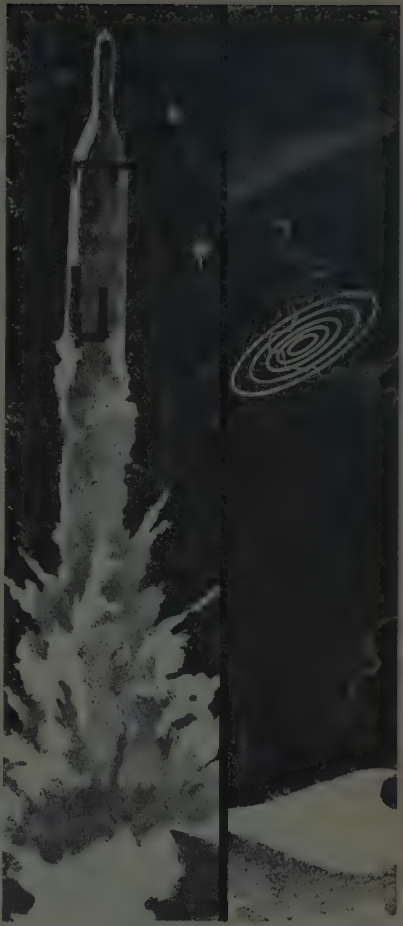
Detailed technical data on the 2N1505 and 2N1506 transistors is available from the factory.

High Power Silicon Rectifier Stack

Pre-engineered high power silicon rectifier stacks incorporating built-in paralleling reactors to insure equal current distribution through parallel circuit branches are now available from **International Rectifier Corp.,** 1521 E. Grand Ave., El Segundo, Calif. The units feature ratings up to 750 amperes, with from 50 to 600 volts peak reverse voltage ratings.

A standard "building block" is a 2-1-2-D "doubler" module with integral parallel-

(Continued on page 167A)



po-si′tion

A NOTE FOR TALENTED ENGINEERS:
We invite your inquiry as to important positions that exist at Nortronics' Precision Products Department (formerly the Military Products Division of American-Standard).

Plan your future with Nortronics' Norwood team and stimulate your professional growth. If you can qualify in one of several electro-mechanical areas you will work on challenging programs with professionally dedicated associates.

FOOTNOTE: *Excellent salaries for qualified engineers in both our Systems and Components Groups.*

Contact E. P. REARDON,
NORTRONICS Precision Products Department,
Norwood, Massachusetts

NORTRONICS
A Division of NORTHROP CORPORATION

Through the looking glass . . .

You'll find new opportunity
new reward
in the new look at Link

It's even more exciting — and certainly more real — than looking-glass land and nearer to now than the closest mirror.

Right now, things are really humming at the Link Division of General Precision, Inc. Link is one company which has accepted the challenge of the years ahead with eager enthusiasm. In fact, projects now under way at Link are among the most advanced in the nation.

At Link, engineers are busily engaged with research and development on "Dialog" systems, sub-systems and components. Another example of Link/ability, "Dialog" systems planning combines digital computation with analog measuring. It offers the best of both—in compatible form—to meet the most exacting control, design, analysis and scheduling specifications.

But let's be honest. Without top caliber engineers we can't really do anything. That's why engineers coming to Link today have Opportunity with a capital "O".

If you are an electronics engineer who thinks big—who has original ideas and follow-through to make them click, why not write today, or send your résumé—in complete confidence—to:

Mr. C. P. Darrah
Link Division
General Precision, Inc.
Binghamton, New York

Mr. R. P. Rutman
Link Division
General Precision, Inc.
Palo Alto, California

LINK DIVISION **GENERAL PRECISION, INC.**

BINGHAMTON, NEW YORK

PALO ALTO, CALIFORNIA • COLLEGE PARK, MARYLAND

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 165A)

ing reactor, and four 70 ampere rated silicon junction rectifiers mounted on copper cooling fins. Two of these modules may be mounted to from a single phase bridge (as shown in photo) rated up to 550 amperes rectified dc output (when operating within recommended temperature and cooling limits).

Three of these basic modules will form a 3-phase bridge rated to 750 amperes. Other configurations include "Scott 4-phase bridges," and 6-phase bridges in both series and parallel connections and proportional ratings.

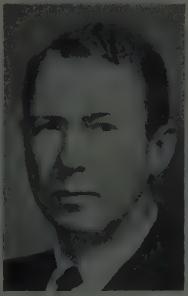
For detailed data on this new high power silicon stack line, request Bulletin SR-336.

Piore and Palmer Appointed By IBM

A separation of the corporate staff research and engineering function of **International Business Machines Corporation** into two functions was announced today by President Thomas J. Watson, Jr.

Dr. Emanuel R. Piore, IBM director of research, will now report directly to A. L Williams, IBM executive vice president. Dr. Piore was also named chairman of the company's Research and Development Board. The board serves to assure the most effective direction and use of IBM's total technical resources.

Ralph L. Palmer, IBM director of engineering, will also now report to Williams and will continue to exercise staff supervision of the company's engineering activities.



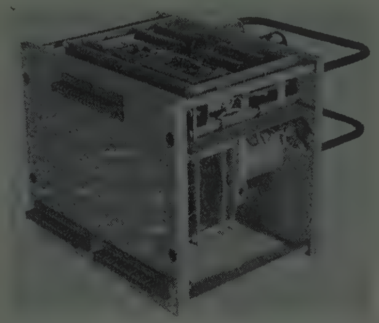
Research and engineering were previously combined under W. W. McDowell, formerly vice president for research and engineering, who was appointed resident vice president for IBM's operations in Endicott, N. Y.

Dr. Piore joined IBM in 1956. He previously was vice president in change of research for the Avco Manufacturing Co. From 1946 to 1955, Dr. Piore was with the Office of Naval Research, becoming ONR's chief scientist. He is a member of President Eisenhower's Science Advisory Committee.

Palmer, an electrical engineering graduate of Union College, Schenectady, N. Y., joined IBM in 1932 at Endicott, N. Y. He subsequently served in various managerial positions in the company's engineering and research operations.

Multiple Circuit Switch

A new, simplified method of switching multiple circuits in television broadcasting operations has been announced by **Telecontrol Corp.**, 1418 W. 166th St., Gardena, Calif.



The new "building block" switching systems design permits simultaneous multiple circuit switching, with a reduced investment in switching equipment. The basic switching modules can be applied to all television switching operations, including studio, master control and distribution switching systems.

Key to the switching module concept is a new type of multiple circuit rotary stepping switch. The remotely controlled switch is adaptable to various methods of control within television broadcast installations of all sizes.

Telecontrol switching modules have a number of inherent advantages over video relays of transistor-diode switching assemblies, according to the firm. Switching is accomplished within a one-inch radius, designed for video switching with extremely low capacitance. This results in low cross-talk and flat frequency response, even when a large number of modules are employed.

The stepping switch can handle simultaneous switching of various combinations of video, sync, audio, control and tally circuits within the same module. It also utilizes a wide range of control flexibility in remote positioning, including pulse control, rotary selector control and pushbutton selection.

Another advantage of the Telecontrol switching concept is that it permits high-speed off-air switching, with rotation to the extreme position in the order of 0.05 seconds.

Distribution switching modules (Series VSA, VSB, VSC) are designed for video distribution exchange systems, house monitoring systems and other applications where off-air pre-selection is permitted.

The distribution modules offer a number of advantages over conventional patching methods, which require a large number of circuit connections, the services of a skilled operator and time for set-up. Remote control switching with the Telecontrol modules sharply reduces the number of required circuit feeds, releases the op-

(Continued on page 168A)

Engineers:
There's new opportunity at Link

Among the opportunities
at Link, Binghamton, are:

DIGITAL COMPUTER ENGINEERS

Several senior staff engineer and senior electronic engineer opportunities have been created by the formation of the new Digital Systems Development Department.

Assignments will include the evaluation of digital systems, covering the responsibility for development project design groups. Important technical contributions will include development work on "NOR" Logic, Direct-Coupled Transistorized Logic, and other techniques for future digital projects.

Requirements include minimum BSEE or BS (Physics) with experience in digital systems design and computer check-out, and/or transistor circuit design.

SENIOR STAFF ENGINEER

To prepare, coordinate and evaluate proposals for contract sponsorship of development programs; assist laboratory manager in evaluating ideas for company sponsored development; investigate new product areas for marketability, investment and present Link capability to government and industrial concerns.

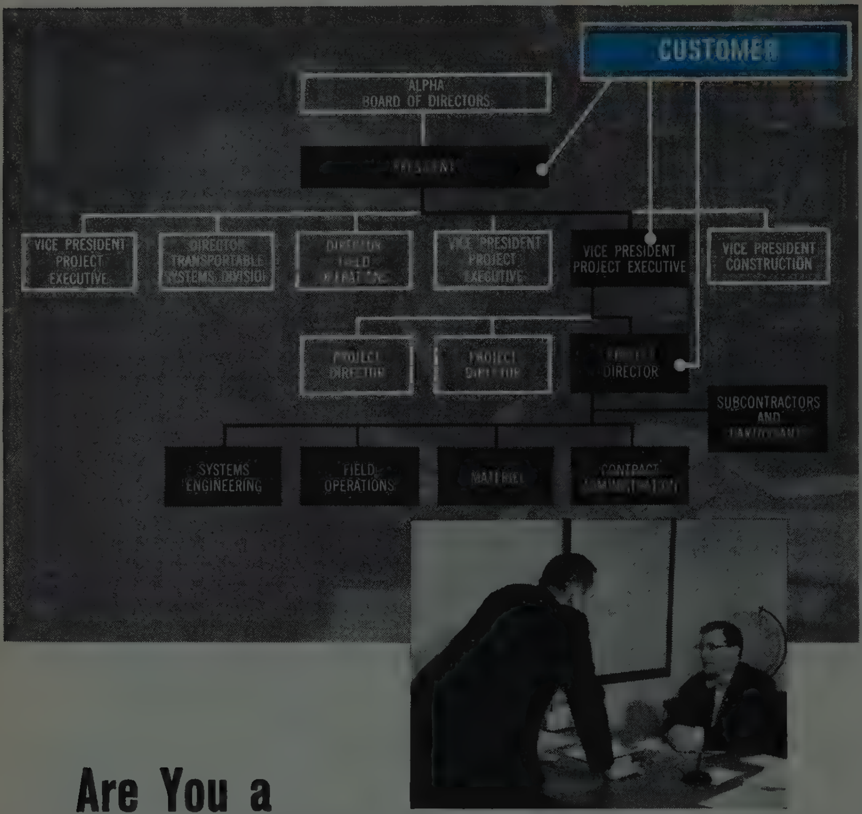
Minimum BSEE with advanced work in communications. MSEE (communications) preferred.

Experience must include 10-12 years in electronic system and hardware design and a broad background in several related areas including: early product design experience in military electronics; design of custom electronic products for commercial aviation.

If you are an electronics engineer with an eye towards a brighter, more rewarding future, why not write or submit your resume—in complete confidence—to Mr. C. P. Darrah.

GP
LINK DIVISION
GENERAL PRECISION, INC.
BINGHAMTON, NEW YORK

alpha is a systems management organization


CUSTOMER
ALPHA BOARD OF DIRECTORS
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VICE PRESIDENT PROJECT EXECUTIVE
DIRECTOR TRANSPORTABLE SYSTEMS DIVISION
DIRECTOR FIELD OPERATIONS
VICE PRESIDENT PROJECT EXECUTIVE
VICE PRESIDENT PROJECT EXECUTIVE
VICE PRESIDENT CONSTRUCTION
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PROJECT DIRECTOR
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SYSTEMS ENGINEERING
FIELD OPERATIONS
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Are You a Systems-oriented Engineer?

Simple and clear-cut, Alpha's internal organization is aligned to provide systems coordination by means of vertical integration of management, engineering, and administration of each project within the division to which the project is assigned. Each project division is under a Vice President and Project Executive who is organizationally a part of the central management of Alpha. Functionally, the Project Director, together with the project's own engineering, operations, materiel, contract administration, and business development personnel, deals directly with the customer and has complete authority, responsibility, and accountability for the timely execution of the project.

Does this team-oriented attitude that engenders effective interfacing relationships and intra-team communications appeal to you? Do you have systems-oriented technical talent with the specialized capabilities needed to determine system parameters; integrate a complex of equipments; direct subsystems suppliers? Write:

C. P. Nelson
Employment Manager

Alpha
CORPORATION
a Subsidiary of Collins Radio Company

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NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

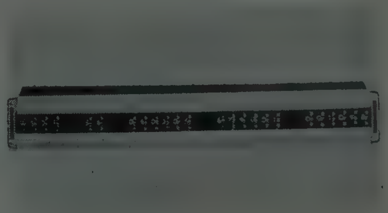
(Continued from page 167A)

erator for other assignments and reduces total setup time to seconds.

Ten or 20 inputs are standard for all distribution and program switching modules. For switching NTSC color video and for large video distribution exchanges, special compensated switching modules are available with built-in equalized impedance and phase shift compensation.

Modular In-Line Switch

The Digitran Co., 660 So. Arroyo Parkway, Pasadena, Calif., announces a new panel switch concept (the new 7300 Series). The Digiswitch replaces ten-position, or octal, switch units of the rotary type. Panel space requirements are reduced so that designers can mount four times as many Digiswitches in the normal space required for rotary switch elements.



The switch is finger controlled and provides for ten-position, single-pole and double-pole, binary coded decimal with variations, and octal 0, 1, 2, 4 coded outputs. Internal lighting of the readout and color coding of the switch tabs are also available.

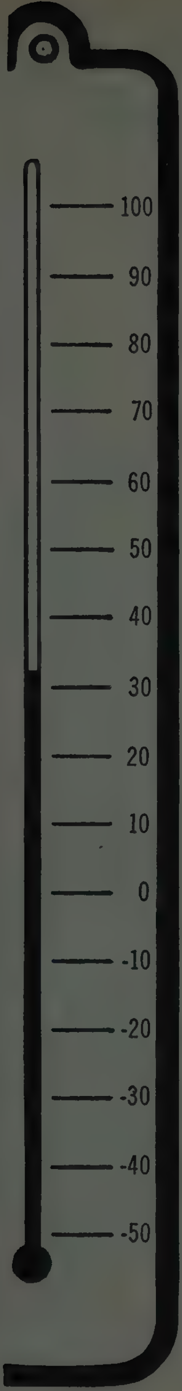
The in-line readout is said to reduce reading errors and simplify switch setting. The unit finds its applications in automation, automatic control systems, computer designs and other systems which require the setting of numerical constants or coefficients.

The Digiswitch is designed in switch modules which permit ganging in any desired number. For descriptive data sheet No. 7300, write to the firm.

Current Sampling Transformer

Valor Instruments, Inc., 13214 Crenshaw Blvd., Gardena, Calif., has introduced the IST series of seven pulse cur-

(Continued on page 170A)

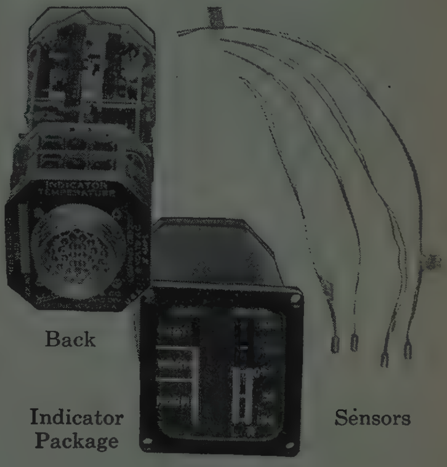


1 or 6 . . .

How do you take the temperature of an electronic system?

Use ONE Compact, Accurate, Lightweight, Completely Transistorized Oster-Developed Temperature Indicating System. Replaces 6 Indicating Systems.

The device illustrated senses and displays the highest of 4 temperature sensor signals on the Convair B-58 supersonic bomber. It also reads out the temperature of the plane's 2 refrigeration systems. However, the unit can monitor temperature of any military or commercial electronic system. Special versions can be furnished to monitor any analog voltage signal.



Back

Indicator Package

Front

Sensors

Oster®

Entire unit (component parts illustrated) has an accuracy of 1½%, weighs only 3.85 lbs. max., requires only 2⅜″ panel space, and does the work of 6 indicating systems. Component parts consist of 4 sensors, a computer package with 6 channels, and a hermetically sealed indicator package.

The unit is typical of Oster designed equipment.

Computer Package

For help with your instrumentation and display problems, talk to the specialists at the Avionic Division.

Burton Browne Advertising

OTHER PRODUCTS INCLUDE

- Servos
- Synchros
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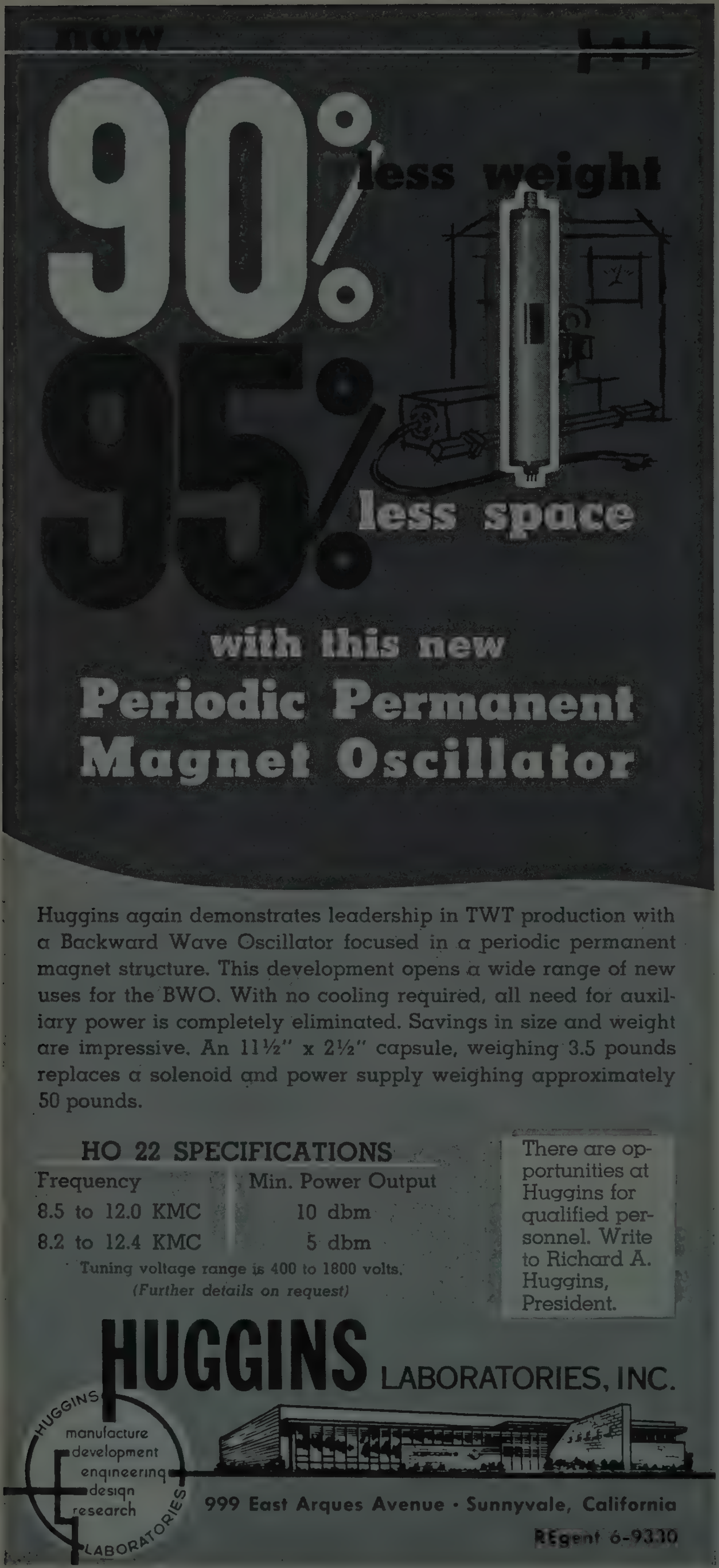
Engineers For Advanced Projects: Interesting, varied work on designing transistor circuits and servo mechanisms.
Contact Mr. Robert Burns, Personnel Manager, in confidence.



Huggins again demonstrates leadership in TWT production with a Backward Wave Oscillator focused in a periodic permanent magnet structure. This development opens a wide range of new uses for the BWO. With no cooling required, all need for auxiliary power is completely eliminated. Savings in size and weight are impressive. An 11½" x 2½" capsule, weighing 3.5 pounds replaces a solenoid and power supply weighing approximately 50 pounds.

HO 22 SPECIFICATIONS

Frequency	Min. Power Output
8.5 to 12.0 KMC	10 dbm
8.2 to 12.4 KMC	5 dbm

Tuning voltage range is 400 to 1800 volts.
(Further details on request)

There are opportunities at Huggins for qualified personnel. Write to Richard A. Huggins, President.

HUGGINS LABORATORIES, INC.

999 East Arques Avenue · Sunnyvale, California
REgent 6-9310

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 168A)

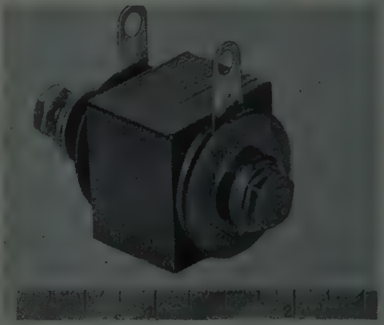
rent sampling transformers which deliver synchronizing voltage pulses for use with radar transmitters or other devices which develop high pulse currents. The voltage pulses have the same shape as the high current pulses. No resistance is added to the circuit because the transformer is not connected to the current carrying the conductor; voltage pulses are developed by passing the conductor through the hole in the transformer. This new approach eliminates the resistive networks used in conventional methods.

Specifications: Size $\frac{9}{16}\times\frac{1}{2}\times\frac{3}{8}''$; Weight $\frac{1}{8}$ ounce; Ratio 20:1 to 150:1; Pulse Widths 0.4 to 3.0 μsec at 50 volts; Inductance 0.12 to 6.0 MH; Optimum Load 50 to 500 ohms; Meets MIL T27A.

Price: 1–4 units $7.70 each; $4.60 each for 100. Delivery: From stock typically; two weeks maximum.

Silicon Rectifiers Surge Protector

Type SP Surge Protectors manufactured by **Electric Products Division, Dept. 502P, Vickers Incorporated,** 1815 Locust St., St. Louis 3, Missouri, protect silicon power rectifiers from breakdown due to transient high voltage.



Non-linear resistance, decreasing with increase in voltage, plus built-in capacitance, absorbs intermittent surge energy up to 3000 watts, limiting voltage to safe value for silicon rectifier. Consumes less than 5 watts under steady-state conditions.

Nine standard types cover the range of 50 to 600 volts normal PIV rating. Field tested for more than a year; laboratory surge tested for more than 5 million cycles.

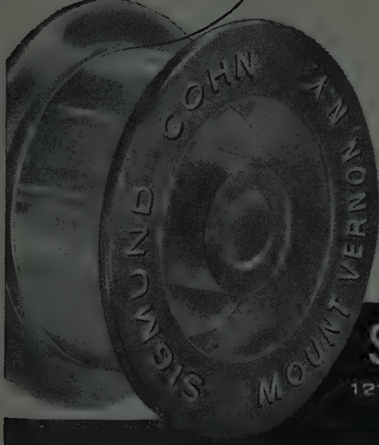
For information write to the firm.

(Continued on page 174A)

Use Your
IRE DIRECTORY!
It's Valuable

Very Small Wire

ELECTROPLATED



Unusual and difficult engineering problems may be solved, at times, by single or multiple electroplating of small wires . . . A wide variety of metals is available, both as the core wire and as the plate. Inquiries invited.

SINCE 1901

Write for list of Products.

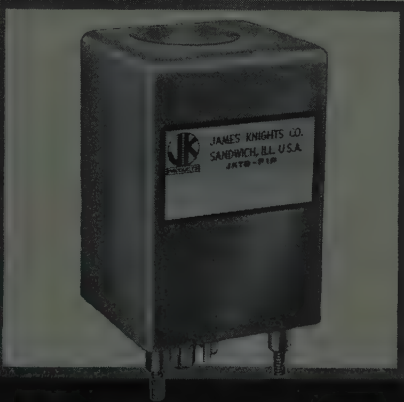
SIGMUND COHN MFG. CO., INC.
121 So. Columbus Ave. Mt. Vernon, N. Y.

1×10^{-8}

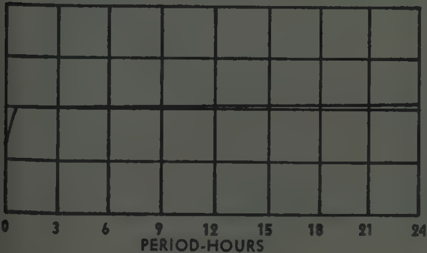
with compact, transistorized series

JKTO-PIP FREQUENCY STANDARDS

(Standard Models Stocked For Quick Delivery)



This series brings together a combination of ultra-stable glass sealed crystals, matched-design silicon transistor-oscillator circuitry, and transistorized proportional control oven performance. The result of this advanced engineering is frequency stability of 1 part in 100,000,000 per day, under normal conditions.



SPECIFICATIONS

FREQUENCY: 1 to 5 mc. is standard. Other frequencies available on special order.
OVEN: Transistorized, proportional control.
OUTPUT: One V into 5000 ohms.
POWER: Operates from 24 to 28V D.C.
DIMENSIONS: 2" x 1.84" x 2⅞" H.
WEIGHT: 10 Oz. Maximum.
ENVIRONMENTAL: Hermetically sealed, shock and vibration resistant.
TEMPERATURE RANGE: From −55°C to +75°C. Also from −55°C to 100+°C.

PRODUCTS

For information, write stating your requirements.
THE JAMES KNIGHTS COMPANY
SANDWICH, ILLINOIS

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a tradition in voltage regulated dc power supplies

Kepco inc.

the standard for the "state of the art" for over 14 years

More than 120 standard active models
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transistorized
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vacuum tube
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magnetic

For complete specifications send for Catalog B-601.

Kepco inc

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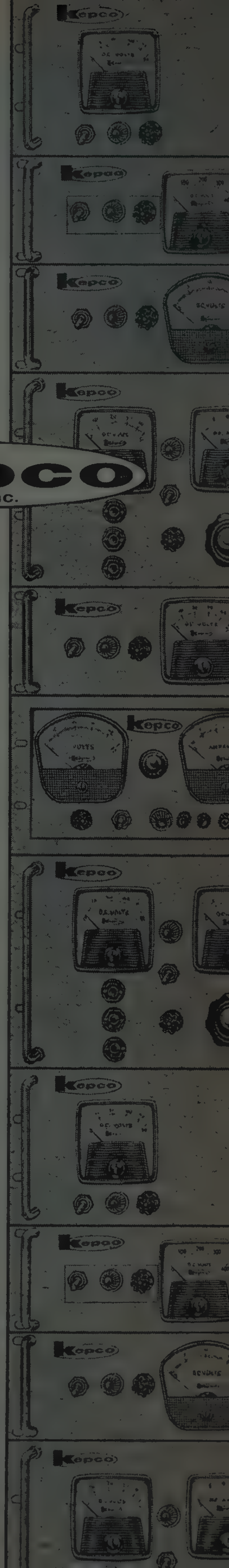


FREQUENCY STANDARDS

PRECISION FORK UNIT
TYPE 50

Size 1" dia. x 3¾" H. Wght., 4 oz.*

Frequencies: 240 to 1000 cycles
Accuracies:—
Type 50 (±.02% at —65° to 85°C)
Type R50 (±.002% at 15° to 35°C)
Double triode and 5 pigtail parts required
Input, Tube heater voltage and B voltage
Output, approx. 5V into 200,000 ohms

*3⅛" high
400 - 1000 cy.

FREQUENCY STANDARD
TYPE 50L

Size 3¾" x 4½" x 5½" High
Weight, 2 lbs.

Frequencies: 50, 60, 75 or 100 cycles
Accuracies:—
Type 50L (±.02% at —65° to 85°C)
Type R50L (±.002% at 15° to 35°C)
Output, 3V into 200,000 ohms
Input, 150 to 300V, B (6V at .6 amps.)

PRECISION FORK UNIT
TYPE 2003

Size 1½" dia. x 4½" H. Wght. 8 oz.*

Frequencies: 200 to 4000 cycles
Accuracies:—
Type 2003 (±.02% at —65° to 85°C)
Type R2003 (±.002% at 15° to 35°C)
Type W2003 (±.005% at —65° to 85°C)
Double triode and 5 pigtail parts required
Input and output same as Type 50, above

*3½" high
400 to 500 cy.
optional

FREQUENCY STANDARD
TYPE 2005

Size, 8" x 8" x 7¼" High
Weight, 14 lbs.

Frequencies: 50 to 400 cycles
(Specify)
Accuracy: ±.001% from 20° to 30°C
Output, 10 Watts at 115 Volts
Input, 115V. (50 to 400 cycles)

FREQUENCY STANDARD
TYPE 2007-6

NEW

TRANSISTORIZED, Silicon Type
Size 1½" dia. x 3½" H. Wght. 7 ozs.

Frequencies: 400 — 500 or 1000 cycles
Accuracies:
2007–6 (± .02% at —50° to +85°C)
R2007–6 (±.002% at +15° to +35°C)
W2007–6 (±.005% at —65° to +125°C)
Input: 10 to 30 Volts, D. C., at 6 ma.
Output: Multitap, 75 to 100,000 ohms

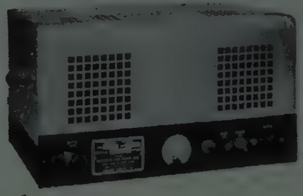
FREQUENCY STANDARD
TYPE 2121A

Size
8¾" x 19" panel
Weight, 25 lbs.
Output: 115V
60 cycles, 10 Watt
Accuracy:
±.001% from 20° to 30°C
Input, 115V (50 to 400 cycles)

FREQUENCY STANDARD
TYPE 2001-2

Size 3¾" x 4½" x 6" H., Wght. 26 oz.

Frequencies: 200 to 3000 cycles
Accuracy: ±.001% at 20° to 30°C
Output: 5V. at 250,000 ohms
Input: Heater voltage, 6.3 - 12 - 28
B voltage, 100 to 300 V., at 5 to 10 ma.



FREQUENCY STANDARD
TYPE 2111C

Size, with cover
10" x 17" x 9" H.
Panel model
10" x 19" x 8¾" H.
Weight, 25 lbs.
Frequencies: 50 to 1000 cycles
Accuracy: (±.002% at 15° to 35°C)
Output: 115V, 75W. Input: 115V, 50 to 75 cycles.

ACCESSORY UNITS
for TYPE 2001-2

L—For low frequencies
multi-vibrator type, 40-200 cy.
D—For low frequencies
counter type, 40-200 cy.
H—For high freqs, up to 20 KC.
M—Power Amplifier, 2W output.
P—Power supply.

This organization makes frequency standards within a range of 30 to 30,000 cycles. They are used extensively by aviation, industry, government departments, armed forces—where maximum accuracy and durability are required.

WHEN REQUESTING INFORMATION
PLEASE SPECIFY TYPE NUMBER

American Time Products, Inc.

Watch Master
Timing Systems

Telephone: PLaza 7-1430

580 Fifth Ave., New York 36, N. Y.

Fits in an 8¾" slot
Beckman/Berkeley
FREQUENCY METER
MODEL 7175
INCHES

Measure 10cps to 110Mc with one compact meter

Comprehensive range for only $1895. Never before has so broad a range been offered for so low a price—a combination made possible by closely integrating a simple heterodyne converter with a top-notch 10Mc counter. Frequencies up to 10Mc are measured by direct counting. To measure frequencies above 10Mc, the operator simply rotates reference frequency selector until panel meter shows strong deflection, then reads counter indication. Measurements take less than a minute to make. Accuracy far exceeds FCC requirements over communications range. Possible error is .00004% or less from 1Mc to 110Mc.

Write for technical bulletin on Model 7175.

Frequency measuring range
10cps to 110Mc
Sensitivity
100mv rms into 1M ohms up to 10Mc
100mv rms into 100 ohms up to 110Mc
Accuracy
Oscillator accuracy ± 1cps
Oscillator stability
3 parts in 10^7 per week
Recording facility
Rear jack carries code signals to actuate Beckman printer
Dimensions:
8¾" x 19" panel, 17" deep
Weight
Ready for rack: approx. 47 lbs.
In cabinet: approx. 60 lbs.
Price $1895

INSTRUMENTAL IN YOUR FUTURE 25 YEARS 1935-1960

Beckman® *Berkeley Division*
Richmond, California

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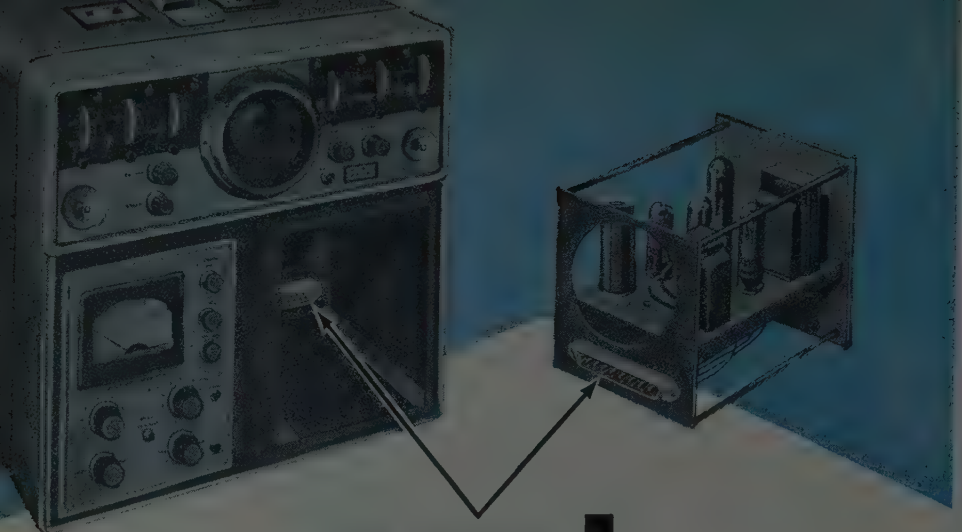


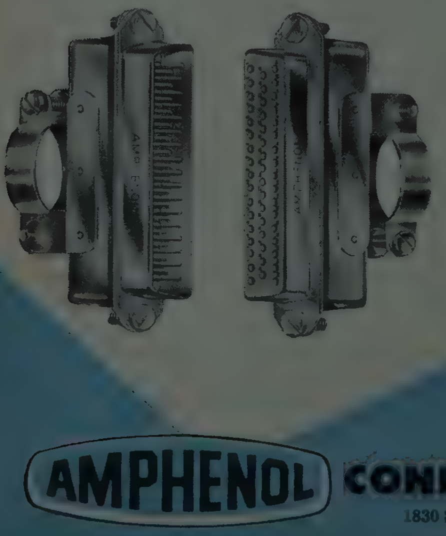
new!

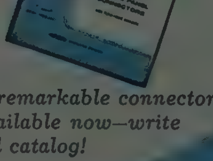
MinRac 17

miniature rack & panel connectors with POKE HOME® contacts

Solve space, weight and size problems with AMPHENOL's new Min Rac 17 connectors, true miniatures with the "Big Plus" advantage of Poke Home contacts! Min Rac 17's are rack & panel connectors ideally suited for today's compact chassis designs, connectors half the size and weight of standards, delivering full size efficiency. And with the patented Poke Home contact concept (U.S. Pat. 2,419,018), Min Rac 17's are easily, reliably assembled—contacts are crimped or soldered outside the connector body, then "poked home" for assembly.

Min Rac 17's are available in 9, 15, 25, 37 and 50 contacts in rack & panel, cable-to-chassis and cable-to-cable designs. Contacts are gold plated. Shells may be ordered with clear chromate or gold iridite finish.





These remarkable connectors are available now—write for full catalog!

AMPHENOL CONNECTOR DIVISION
1830 S. 54th AVE., CHICAGO 50, ILLINOIS
Amphenol-Borg Electronics Corporation

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 170A)

Peaked Amplifier

CES Electronic Products, Inc., P.O. Box 7504, San Diego 7, Calif., has announced development and production of a new peaked amplifier in the 15 to 2000 cps range.

The new unit, Model 107-B, is a three stage direct-coupled amplifier with a CES Twin-T Filter in the negative feedback loop. The filter gives at least 50 db insertion loss at its balance frequency and the normal feedback is at least 70 db. This retains 20 db feedback for stabilization at the peaked frequency and results in a very sharp, deep-shouldered response curve. The performance of the amplifier is independent of the driving point impedance and can be terminated in any high impedance load. Q's of 100 are attainable with excellent stability.

According to the company, the amplifier is available in standard models for frequencies from 15 to 2000 cps. Special models with frequencies as low as 0.01 cps can be supplied on special order.

For additional information, write to the firm.

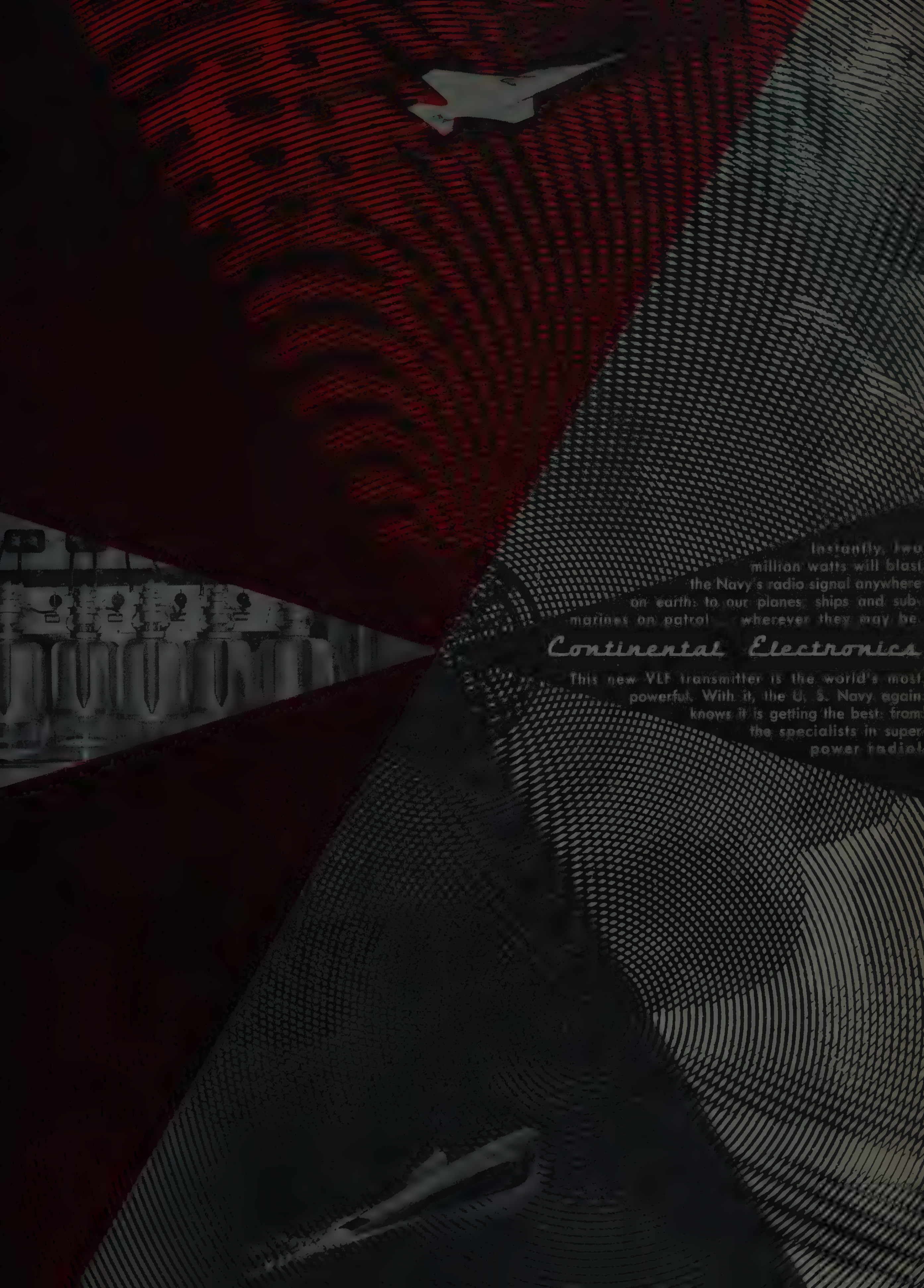
Rotary Solenoids And Stepping Motors

A new application guide to Ledex rotary solenoids, selectors and Syncramental stepping motors has just been issued by **G. H. Leland, Inc.,** 123 Webster St., Dayton, Ohio.

This eight-page bulletin (No. A-1259) —illustrated with line drawings, charts, photographs and circuit diagrams—contains specific physical, performance and environmental data on more than 250 different models available for immediate delivery. It also supplies information on how to use this data to select the proper rotary solenoid, selector or Syncramental stepping motor for any application.

Only half the size of ordinary units, Leland's highly efficient Ledex rotary solenoid is capable of developing twice as much power as these ordinary solenoids. Its stepping and selector switch—

(Continued on page 176A)



Instantly, two
million watts will blast
the Navy's radio signal anywhere
on earth to our planes, ships and sub-
marines on patrol . . . wherever they may be.

Continental Electronics

This new VLF transmitter is the world's most
powerful. With it, the U. S. Navy again
knows it is getting the best from
the specialists in super
power radio

Give your products
MORE RELIABILITY and
BETTER PERFORMANCE with
FREED
QUALITY
AVAILABLE FROM STOCK

MINIATURE AUDIO TRANSFORMERS to MIL Specifications

- Hermetically sealed Grade 4.
- Frequency response:
  ± 2DB 30-20,000 CPS
  *± 2DB 200-10,000 CPS

Cat. No.	Imped. level —	Appl.	MIL Type
PMA 1	Pri. 50/200/500 Sec. 60,000 C.T.	Line or mike to single or P.P. grids	TF4RX10YY
PMA 2	Pri. 4/8 Sec. 60,000 C.T.	Dynamic mike or spkr. voice coil to single or P.P. grids	TF4RX10YY
PMA 3	Pri. 50/200/500 Sec. 60,000 C.T.	Line or mike to single or P.P. grids	TF4RX10YY
PMA 4	Pri. 15,000 Sec. 60,000 C.T.	Single triode plate to single or P.P. grids	TF4RX15YY
PMA 5*	Pri. 15,000 Sec. 60,000 C.T.	Single triode plate to P.P. grids	TF4RX15YY
PMA 6	Pri. 15,000 Sec. 50/200/500	Single triode plate to multiple line	TF4RX13YY
PMA 7*	Pri. 15,000 Sec. 50/200/500	Single triode plate to multiple line	TF4RX13YY
PMA 8	Pri 30,000 C.T. Sec. 50/200/500	Push-pull triode plate to multiple line	TF4RX13YY
PMA 9	Pri. 60,000 C.T. Sec. 50/200/500	Crystal mike or pickup to multiple line	TF4RX13YY
PMA 10	Pri. 50/200 Sec. 50/200/500	Mixing or matching	TF4RX16YY
PMA 11	40 hy, 3 ma. d.c. 3500 Ω d.c. res.	Parallel feed reactor	TF4RX20YY

MAGNETIC AMPLIFIERS
- Hermetically Sealed To MIL Specifications
- No Tubes
- Direct Operation from Line Voltage
- Fast Response
- Long Life Trouble Free Operation
- Phase Reversible Output

Power Gain 2 x 10⁸

Transistor Preamp. MAT-1 Wt. 10 oz.
Mag. Amp. MAF-5 Wt. 18 oz.
Motor

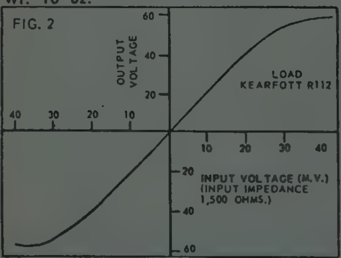


FIG. 2

WRITE FOR FURTHER INFORMATION ON THESE UNITS OR SPECIAL DESIGNS.
Partial listing only—Send for NEW 48 page transformer catalog. Also ask for complete laboratory test instrument catalog.

FREED TRANSFORMER CO., INC.
1720 Weirfield St., Brooklyn (Ridgewood) 27, N.Y.

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Contnued from page 174A)

which can be manufactured in hermetically-sealed containers—has replaced the use of numerous ordinary relays in many applications where space and weight present a problem. Ledex stepping and selector switches require only one-fifth the space formerly taken up by ordinary relays.

A versatile, high-precision, bi-directional stepping motor, the Ledex Syncramental motor develops at least three times greater torque than comparable devices can provide. During its design life of 2,000,000 steps in each direction, the probability of its missing a step is only 1 in 1,000,000.

600-Channel Transistorized Multiplex Carrier

General Electric Co., Communication Products Dept., Lynchburg, Va., announced today that it has developed a transistorized multiplex-carrier system capable of handling up to 600 voice frequency channels on a single radio beam.

Fully transistorized, the new General Electric single sideband suppressed carrier system has toll quality capable of meeting international and domestic long-distance standards.

In the new multiplex-carrier, reliability is enhanced by the use of a technique which utilizes a standby component for each component in the common equipment, such as amplifiers and master oscillators. Upon failure of an individual part, the parallel component keeps operating, with no switching delay to seriously affect data transmission."

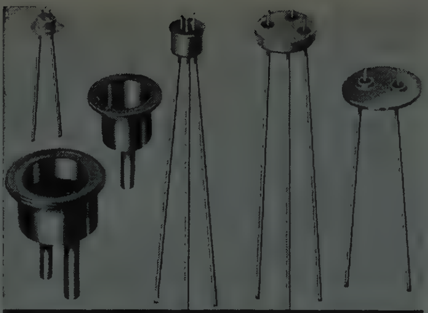
Also achieved is a substantial reduction in power drain. Typical equipment currently available requires thousands of watts for a system utilizing 240 channels. This design takes less than 700 watts for the corresponding capacity.

Despite the increase in capacity which makes 600 channels available, this unit is one-third the size of conventional tubed equipment. G-E's design makes possible 120 channels in an eight-foot space where only 24 channels previously could be accommodated.

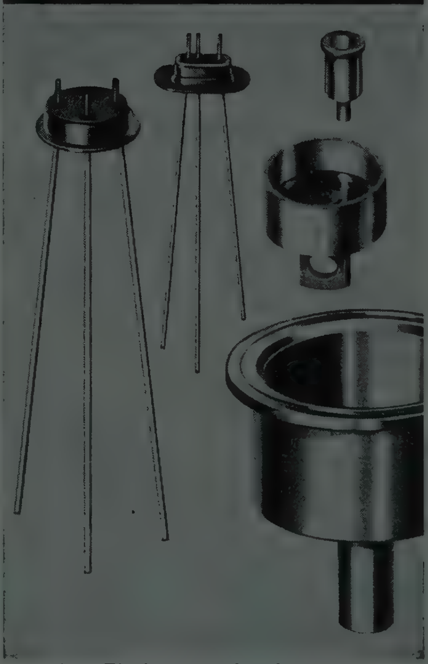
Computer Grade Electrolytics

A new four-page catalog describing Type QE computer grade electrolytic capacitors is available from **Aerovox Corp.,**

(Continued on page 178A)



Quality
Is The Constant In
GLASS-TITE
HERMETIC SEALS



Glass-Tite's story is simple and successful. They make a custom seal to your most minute specifications or they supply standard glass-to-metal seals — either way they make only one quality . . . THE BEST. No design requirement is too tough, no reasonable delivery date is impossible.

Specify Glass-Tite in Kovar and compression seals, diodes and rectifier enclosures for the semi-conductor industry and all electronic applications.

Write Dept. 729 for literature and send details of your design requirements.

[illegible]

GLASS-TITE
INDUSTRIES, INC.
GT
725 BRANCH AVENUE
PROVIDENCE, R. I.

Spincasting

A significant contribution to Antenna Technology by D. S. KENNEDY & CO.

Kennedy engineers and scientists have successfully applied a simple scientific principle to the manufacture of large antennas. The application of this principle that "the surface of a liquid when spun about a vertical axis assumes a paraboloidal shape" will result in several important benefits.

First, surface accuracy of antennas should improve by more than an order of ten. This is possible even though the surface to which the liquid is spuncast is irregular, because such irregularities in the container in no way affect the smooth accuracy of the surface of the liquid.

This means substantial savings in time and money in the manufacture of antenna dishes of this accuracy, as well as antennas of greater efficiency.

There is no theoretical limit to the size of dish Kennedy may be able to Spincast. Dishes up to 10 feet in diameter are already in production and Kennedy is ready to receive inquiries as to the application of this technique to your problems. For complete information write, wire or call collect — EVergreen 3-1200.

D. S. KENNEDY & CO.

TRACKING ANTENNAS • RADIO TELESCOPES
RADAR ANTENNAS • SCATTER COMMUNICATIONS ANTENNAS
MICROWAVE RESEARCH AND DEVELOPMENT
TRANSMISSION TOWERS • SWITCHYARD STRUCTURES

Antenna Division
Cohasset, Massachusetts
EVergreen 3-1200

Anchor Metals Division
Hurst, Texas
(Fort Worth) ATlas 4-2583

Advanced Engineering Division
Monterey, California
FRontier 3-2461

AP-710-1
INPUT TUNING
INPUT
OUTPUT TUNING
OUTPUT

SAVE VALUABLE COMMUNICATION DOLLARS with RIXON'S LOW NOISE PREAMPLIFIERS

Much effort has been made and is being expended to reduce "noise" in receiving systems because of the potential dollar savings resulting from improved receiver performance. Even if the "noise" does not materially effect the performance of the receiving system, if there is a long transmission line between the antenna and the receiver, the signal power is reduced by the transmission loss in the line.

The Rixon 400-500 mc preamplifier, shown here, with noise figures in the 4 to 5 db-range, makes a major improvement when placed in front of a 10 to 12-db commercial receiver. It's the same as a four time increase in transmitter power.

Other Rixon Preamplifier Models are available on short notice and are easily adapted at the factory to cover any band in the 10-500 mc frequency range.

SPECIFICATIONS

AP-710-1 UHF

Frequency Range	400 to 500 mc
Noise Figure	3.5 db to 5.5 db*
Gain & Bandwith	21 to 23 db—2 mc
Input & Output Impedance	50 ohms
Max. Input Level	100 mv
Panel Size & Wt.	3½ of 19-in. Rack Space—9¾ lb.
RF Connectors	Type 'N'
Power	115 vac, 50-60 cps, 15 watts
Tube Type	7077 Ceramic Triode
Circuit	Grounded-grid, 2 stages
Tuned Circuits	Three continuously tunable coaxial cavities of rigid brass tubing. Simple peaking adjustment only.

*Depending on tube selection

RIXON ELECTRONICS, INC.
2414 Reedie Dr., Silver Spring, Md. • LO 5-4578

tubes and semi-conductors.

Over 5000 types always in stock at E. T. S.,

world's largest tube and transistor specialist. Special purpose, Transmitting, Receiving. All important brands. Name the type, name the brand, name the quantity—we have it. Famous nationwide service—fast delivery anywhere. *Get your free subscription to the E.T.S. Bulletins* for authoritative, up-to-date new-item and new-availability news—write on your company letterhead today.

ELECTRONIC TUBE SALES INC.
74 Cortlandt St., New York 7, N. Y. BArclay 7-4140 TWX-NY1-4042

H LAB

Model 802B Twin Transistorized Supply

EACH OUTPUT:

0-36 VOLTS

0-1.5 AMPERES

PRICE: $580.00

LINE REGULATION: Less than 5.0 millivolts
LOAD REGULATION: Less than 5.0 millivolts
RIPPLE AND NOISE: Less than 200 μv rms
SERIES CONNECTED: 0-72 volts 0-1.5 amperes

OUTPUT CONTINUOUSLY VARIABLE
REMOTE ERROR SENSING
AUTOMATIC OVERLOAD PROTECTION
CONVECTION COOLING. No moving parts

HARRISON LABORATORIES, INC.
45 INDUSTRIAL ROAD • BERKELEY HEIGHTS, NEW JERSEY • CR 3-9123

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 176A)

New Bedford, Mass. Literature provides complete technical information on these high reliability units including dimensional drawings, performance characteristics and table of stock values. Copies may be had by writing to the firm for QE Bulletin NPJ-110.

DC Amplifier

Packard Bell Computer Corp., a Division of Packard Bell Electronic Corp., 12333 West Olympic Blvd., Los Angeles 64, Calif., announces a completely solid-state DC amplifier, designated the DCA-1-B, which has the modularized plug-in feature illustrated in the photograph. This feature combined with the small size ($4\frac{3}{4}''\times3\frac{1}{2}''\times4\frac{1}{4}''$) permits greater design flexibility as it enables the design engineer to incorporate the amplifier in either a standard or nonstandard case.

The DCA-1-B is an operational, wide band, high accuracy amplifier with a solid-state chopper. Solid state design and the use of a non-mechanical chopper are said to make this a reliable instrument.

Specifications include: Frequency response exceeding −3 db at 200 kc; gain accuracy of ±0.03%; linearity and stability at ±0.02%; and drift of 500 μv referred to input.

Photoelectric Pulse Generator

A shaft-driven device that delivers electrical pulses at its output terminals is available from **W. & L. E. Gurley, Industrial**

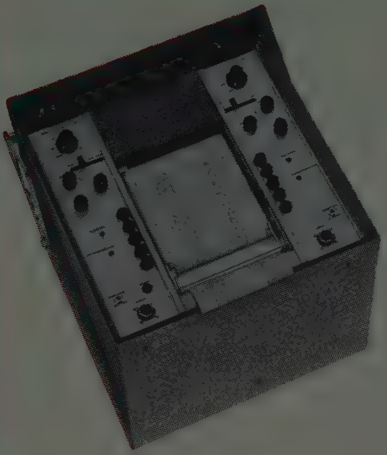
(Continued on page 180A)

Use Your IRE DIRECTORY! It's Valuable

keep an accurate graphic record

OF RESEARCH, DESIGN, TEST DATA

two channels



For General Purpose DC Recording — Model 320

For recording *two variables* simultaneously, the Model 320 provides a versatile, transistorized amplifier for each input signal. The rugged 2-channel recorder assembly has heated stylus recording on two *50 mm wide* rectangular coordinate channels, 4 pushbutton chart speeds, and 6 inches of visible chart. The Recorder can be placed vertically, horizontally or at a 20° angle.

MODEL 320 SPECIFICATIONS

Sensitivity: 0.5, 1, 2, 5, 10, 20 mv/mm and v/cm

Frequency Response: 3 db down at 125 cps, 10 mm peak-to-peak

Common Mode Voltage: ±500 volts max.

Common Mode Rejection: 140 db min. DC

Calibration: 10 mv internal ±1%

Output Connectors for each channel accept external monitoring 'scope or meter

NEW SANBORN PORTABLE DIRECT WRITING RECORDERS FOR IN-PLANT, LABORATORY OR FIELD RECORDING

Two models of this 21 lb. brief case size recorder are available — Model 301 for AC strain gage recording, Model 299 for general purpose DC recording. Both provide immediately visible, inkless traces by heated stylus on 40 division rectangular coordinate charts . . . frequency response to 100 cps . . . 5 and 50 mm/sec chart speeds . . . approx. 4 inches of record visible in top panel window.

single channel

MODEL 301 SPECIFICATIONS

The amplifier section of the Model 301 is an all-transistorized carrier type with phase sensitive demodulator. The power supply and internal oscillator circuits are also transistorized.

Sensitivity: 10 uv rms/div (from transducer)

Attenuator Ratios: 2, 5, 10, 20, 50, 100, 200

Carrier Frequency: 2400 cps internal

Transducer Impedance: 100 ohms min.

Calibration: 40 uv/volt of excitation

Output Connector: for external monitoring 'scope or meter

MODEL 299 SPECIFICATIONS

Combines the dependability of transistors with the high input impedance of vacuum tubes for reliable broad-band DC recording.

Sensitivity: 10, 20, 50, 100, 200, 500 mv/div and 1, 2, 5 and 10 v/div

Input Resistance: 5 megohms balanced each side to ground

Common Mode Voltage: ±2.5 volts max. at 10 mv/div sensitivity increasing to ±500 volts max. at other sensitivities

Common Mode Rejection: 50:1 most sensitive range

Calibration: 0.2 volt internal ±1%

Output Connector: for external monitoring 'scope or meter

Contact your Sanborn Sales-Engineering representative for complete information, or write the main office in Waltham. Sales-Engineering representatives are located in principal cities throughout the United States, Canada and foreign countries.

SANBORN COMPANY

INDUSTRIAL DIVISION

175 Wyman Street, Waltham 54, Mass.

I want a "Whatcha-ma-call-it?"
It's a "thinga-ma-jig that ......"

How often have you struggled to remember the name of a component or electronic item. . . . Just could not think quickly what it is called?

YOU CAN FIND IT IN THE IRE DIRECTORY!

because

(1) The IRE Directory classifies products by purpose and use.
(2) Its listings are fundamental—the way an engineer thinks.
(3) "Terminology" is cross indexed in the pink pages—condensed, simple, not mixed in with firm names.
(4) Ads face listings thus helping to identify products by actual pictures.
(5) Product code numbers reduce complex and duplicate listings, saving you "searching" time and effort.

Study it . . . save it . . . work it.

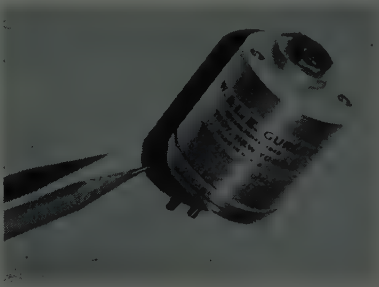
IRE ®

THE INSTITUTE OF RADIO ENGINEERS
1 East 79th Street, New York 21, N.Y.

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 178A)

Division, Troy, N. Y. The Model 8601 Photoelectric Pulse Generator may be used either as a rate generator or as an angle-measuring unit.



When used as a rate generator, the output frequency is read in terms of shaft speed. When used to measure angles, the total angle is determined by totalizing individual pulses. The pulses are generated by a light which passes to a photocell through rotating disc with alternating clear and opaque sectors. The number of pulses per revolution, therefore, is a function of the number of segments on the disc.

The Gurley Generator can be supplied

(Continued on page 183A)

How to Step Up
PRODUCTION and
LOWER the COST of

COIL
WINDING

SHOW ROOM
114 Liberty St.
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Demonstration by
Appointment or
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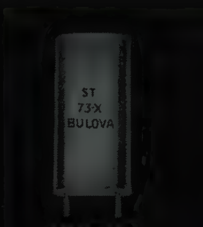
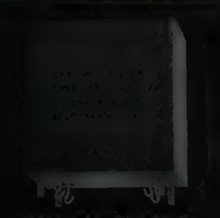
WHitehall 3-1754



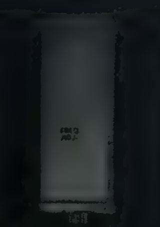
CRYSTALS
Available for any frequency from 1 KC thru 100 mc, to meet and exceed all military specifications.



CRYSTAL FILTERS
Center frequencies from 10 KC thru 20 mc. Band widths of .01% to 8% of center frequency.

BULOVA WATCH CO

OVENS
Complete from multi-purpose AM-100 to the specific BHC series, temperature stabilization to .1°C



PACKAGED OSCILLATORS
10 cps thru 70 mc frequencies available in tube type or transistorized oscillators.

BULOVA

NUMBER ONE SOURCE

in frequency control—components and systems

Today, an increasing number of electronic systems demand a degree of reliability heretofore unobtainable. That is why more and more manufacturers are specifying Bulova.

Bulova crystals, filters, ovens, packaged oscillators, and Bulova frequency control systems, custom-designed for either limited or mass production, meet and exceed military and industrial specifications.

Bulova's experience in mastering many of the most difficult problems involving component and system reliability has made it the *number one source* for frequency control devices. This experience can prove of immense value in your particular program. For more information write Dept. A-1183, today.

BULOVA

ELECTRONICS DIVISION • WOODSIDE 77, NEW YORK

LING'S LIGHT-ARMATURE SHAKER LETS YOU TEST UNDER PRESSURE

Altitude, temperature, humidity! To help you conduct vibration tests under such extremes, Ling brings you the new liquid-cooled Shaker A246 of revolutionary design. In this 7500 force-pound shaker, Ling shaves the weight of the armature to a new low of only 68 pounds—the lightest armature by far for this force-rating. Structural resonance, first major, develops at 2570 cps, bare table. Shaker efficiency is at a new high; the A246 delivers full output at reduced amplifier power, cutting costs on associated electronics equipment. In a chamber, it functions at extremes well above ordinary shakers—from −100°F to +300°F, and up to 125,000 ft. Further, it simplifies chamber testing even more when used with the piggy-back chamber shown above. With Ling seals and baffles, the shaker body acts as one wall of the chamber, and only the table rides into the chamber. This is just one more advance from Ling research; for electronics that always help you out of prototype into production, *fast*—look to Ling. For details, write Dept. IRE-1, at either address below.

LING
ELECTRONICS

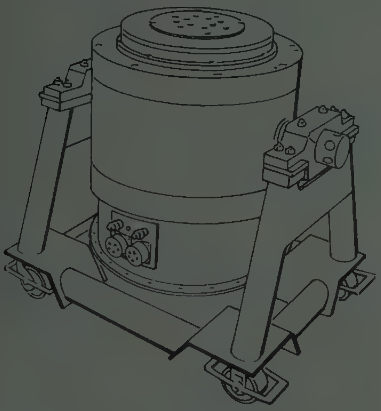
A DIVISION OF LING-ALTEC ELECTRONICS, INC. • 1515 SOUTH MANCHESTER, ANAHEIM, CALIFORNIA • 120 CROSS STREET, WINCHESTER, MASSACHUSETTS

LING ELECTRONICS

The new A246 shaker is but one of the advances growing out of Ling Electronics' continuing program of research and development. This program contributes to the many advantages enjoyed by the Ling customer—fast deliveries, sound engineering and design, ease of maintenance, and the compatibility of Ling environmental testing equipment with other systems.

The compatible design of the A246, for example, permits it to function as part of a test chamber—reducing the size of a chamber needed, and eliminating usual more costly installations. For this method, Ling also supplies a complete line of thermal barriers needed for piggy-back mounting, making combined-environment testing more practical.

Whatever your needs in high-power electronics—for vibration testing, acoustics or sonar—rely on Ling for truly practical design and advanced engineering.



Model A246 SHAKER, which is illustrated above, offers these other performance advantages: 7500 force pound rating, with high first resonance of 2750 cps. Engineered to operate continuously at maximum force on low input. Features simplified compensation over wide bandwidths, dual magnetic field structure for low stray field and improved force-current linearity.

LING ELECTRONICS • HIGH-POWER ELECTRONICS FOR: VIBRATION TESTING • ACOUSTICS • SONAR

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

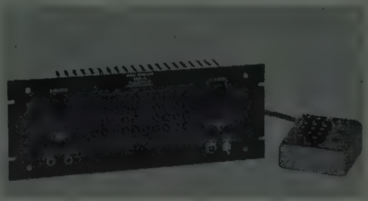
(Continued from page 180A)

with either photoresistive or photovoltaic cell. Output of the former cell consists of alternations between two levels of internal resistance; while output of the latter is a series of dc voltage pulses. Typically the output voltage is 0.1 to 0.25 v at 50K load impedance. Several standard disc patterns provide up to 500 pulses per revolution.

An illustrated bulletin is available.

Digital Translator

A new digital translator which accepts virtually any kind of digital data and provides an analog output plus control signals is now available from **F. L. Moseley Co.,** 409 N. Fair Oaks Ave., Pasadena, Calif.



The instrument, Model 42, permits automatic operation of Moseley Autograf or similar X-Y recorders. Accuracy of the digital-to-analog conversion is 0.1% and the accuracy of the recorders is maintained.

Model 42 is compatible with IBM summary punches and card readers including Models 514, 519, 523, 524, 526, etc. It may also be driven by mechanical punched tape readers such as Friden, Soroban solenoid and Teletype motorized readers, without modification to either the translator or driving equipment.

The new digital translator is supplied with a 10-key serial keyboard for manual input. The instrument accepts 4 digits and sine per axis and provides a front panel display of matrix contents.

Accessories available include a magnetic tape adapter, Flexowriter converter, remote decimal contents read-out panel and an optical magnetic tape converter.

Further information may be obtained from the firm.

Machine Tool Control Brochure Available

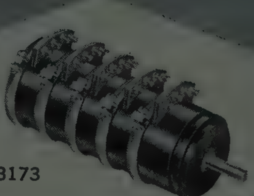
A new brochure describing numerical machine tool controls is now available from the **Industrial Systems Div., Hughes Aircraft Co.,** International Airport Station, Los Angeles 45, Calif.

The new control provides automatic positioning of a two-axis table from a preprogrammed punched one-inch tape. A third axis control can be added to the system.

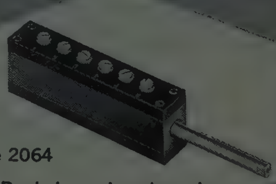
The booklet gives a complete description of the unit, which can be applied to

(Continued on page 194A)

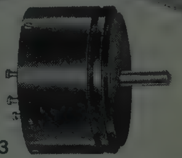
MARKITE
CONDUCTIVE PLASTIC POTENTIOMETERS



Type 3173
⅞" dia. rotary



Type 2064
Dual-element rectangular rectilinear



Type 3033
$1\frac{5}{16}$" dia. rotary

When the ultimate in quality and reliability is required . . . when there is no time for standby or interruptions . . . no room for component value variations . . . no tolerance of failure — then it's high time to specify MARKITE precision potentiometers. Here are only a few reasons why they provide performance beyond the expected:

• Linear stability for more than 50 million cycles • Substantially infinite resolution • Independent linearity to 0.05% in $1\frac{5}{16}$" dia. units and 0.01% in 5" dia. units • Operation in ambient temperatures up to 200° C • Shock and acceleration resistance in excess of 100g • Rotational speeds up to 1,000 rpm • Meet Military Specifications.

Write for Design Data and Catalog for Rotary and Rectilinear Potentiometers.

MARKITE CORPORATION
155 Waverly Place • New York 14, N. Y.

IMPULSE

A DIGEST OF NEW DEVELOPMENTS IN ELECTRONICS AND AUTOMATION

PUBLISHED BY ROME CABLE DIV. OF ALCOA, ROME, N. Y.
PIONEERS IN INSTRUMENTATION CABLE ENGINEERING

STILL A MAN'S WORLD. More attention by designers of electronic equipment to the human factor is being called for by the people handling human engineering for the Air Force. Both operation and maintenance are involved. Typical deficiencies that may unnecessarily complicate maintenance, for example, are: insufficient or inadequate check points, extensive calibration requirements and the placing of high-failure-rate components in inaccessible spots. Ultimate goal is equipment designed so that required maintenance can be conveniently handled by those with a minimum of training in electronics.

UP TWENTY PER CENT. Though not completely clarified, the Defense Department's budget for 1961 in terms of military electronics looks like it will hit $5.5 billion. This is an increase of some 20 per cent over the 1960 figure.

SUN NEVER SETS. True to the traditions of the Empire yesteryear, the British will soon find another means of linking the members of the Commonwealth. This time it will be an undersea cable network that can carry slow-scan TV as well as telephony. The first link, between Britain and Canada, is scheduled for completion in 1961. By 1964 the second leg, tying Canada to Australia, will be ready. And so on.

TERA, GIGA, NANO AND PICO. These are the four prefixes that the National Bureau of Standards has picked following recommendation of the International Committee on Weights and Measures. These four join the already-in-use: mega, kilo, hecto, deka, deci, centi, milli and micro. Tera is 10^{12}; giga, 10^{9}; nano, 10^{-9}; and pico, 10^{-12}. Symbols for these new prefixes are, respectively, T, G, n and p. For example, 10^{-12} farad is a picofarad and is written in abbreviated form thusly: 1 pf.

JURY-SIZED SUB? The Navy is deep into a program (along with nine private firms and research groups) to put into the brine a fully automated sub that will need a crew of only 12. Automated and integrated will be the five fundamental control areas, or "loops": ship control, communications, engineering, weapons and environmental.

CABLEMAN'S CORNER. The old adage "Don't put the cart before the horse" was never so true as it is in these days of automation and instrumentation. With all the intricate pieces of equipment being designed these days, it is important that careful consideration be given to the wire and cable that may be employed in any system. Often forgotten is the unromantic aspect of the connecting links of the system. Cables are the arteries through which must flow the power and informational pulses necessary for reliable performance.

Don't take a chance on being able to obtain a cable that will fit into what is left. Many times, important characteristics such as conductor size, insulating walls, protective sheaths, flexibility and flex-life have to be sacrificed. Don't sacrifice reliability in your cables for an existing space or connector fittings.

For 100% reliability in multi-conductor cables, call on a cable specialist—and call on him as soon as possible. Phone Rome 3000, or write: Rome Cable Division of Alcoa, Dept. 1240, Rome, New York.

These news items represent a digest of information found in many of the publications and periodicals of the electronics industry or related industries. They appear in brief here for easy and concentrated reading. Further information on each can be found in the original source material. Sources will be forwarded on request. ⑰

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EMI

E.M.I. klystrons and magnetrons are available in a wide range of types and specifications to meet your particular and exacting requirements ■ Through the Hoffman Electron Tube Corp. — U.S. representative for E.M.I. — these tubes can be promptly delivered from inventory. Rugged, reliable E.M.I. klystrons and cavities range from 0.5 to 46.0 KMC . . . Magnetrons are available as packaged types for pulse operation; in "E", "Q", and "J" bands; forced-air cooled; indirectly heated cathodes; millimeter operation ■ E.M.I. Electronics Ltd. is one of the world's leading pioneers in the development of special and standard high-performance tubes for military and commercial use.

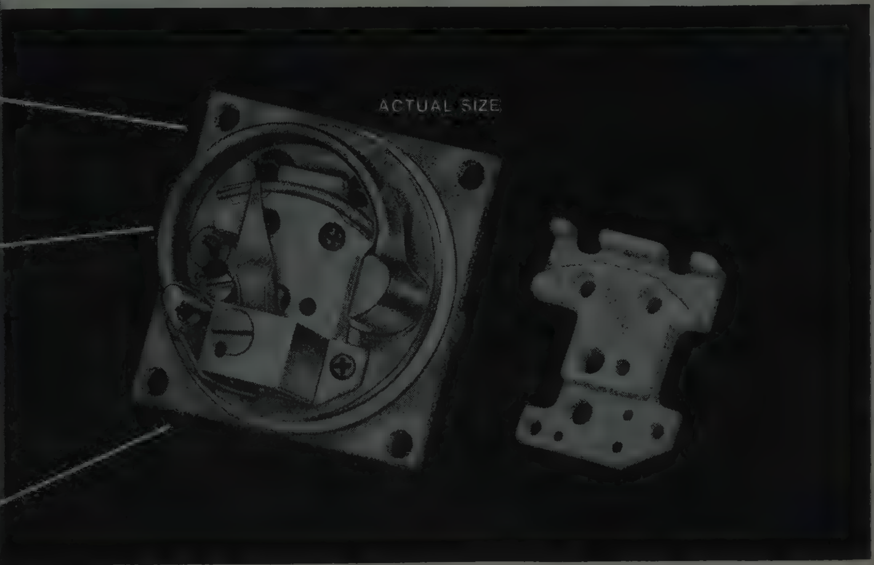
You can rely on E.M.I. for the highest standard of accuracy and reliability.

Send for your copy of the new E.M.I. 300-page catalog.

Hoffman Electron Tube Corp.
804 Newbridge Avenue
Westbury, New York

SUPRAMICA® 555 ceramoplastic

the world's most nearly perfect precision-moldable electronic insulation



for total reliability... at high temperatures ...specified in BOURNS Inc. transducers

Why did BOURNS, INC. select SUPRAMICA 555 ceramoplastic as the insulating base for its ultra high-temperature differential pressure transducers?

BOURNS' engineers cite three reasons ... each a contribution to the total reliability of these airborne telemetering devices. "First is temperature. The sensitive element of the mechanism must withstand high operating temperatures. Next, SUPRAMICA 555 offers a combination of excellent insulating characteristics, which are essential to the highly accurate functioning of the potentiometer. In addition, this ceramoplastic material is readily moldable into complex shapes, such as that required for this intricate part."

For other applications SUPRAMICA 555 is used under operating conditions as high as +700°F. ... SUPRAMICA 555 is one of the many ceramoplastic and glass-bonded mica insulation materials produced by MYCALEX CORPORATION OF AMERICA, in precision-molded and machinable formulations. Whatever your insulation need there is a MYCALEX product to meet it—for example, SUPRAMICA 620 machinable ceramoplastic, which has a maximum operating temperature of +1550°F. Write today outlining your design problem for specific information.

THE FIRST SYNTHAMICA® SYNTHETIC MICA

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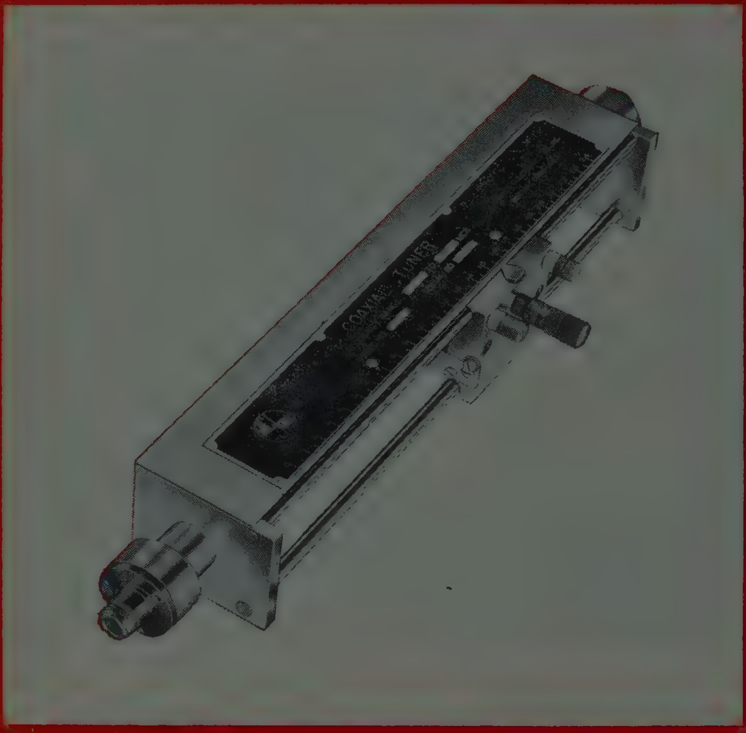
WORLD'S LARGEST MANUFACTURER OF GLASS-BONDED MICA AND CERAMOPLASTIC PRODUCTS

PRECISE *MicroMatch*®

COAXIAL TUNERS

TUNE TO

VSWR 1.000 200-4000 MCS.



MAKES YOUR LOAD A REFLECTIONLESS TERMINATION

DESIGNED FOR USE whenever extremely accurate RF power terminations are required. This laboratory type Coaxial Tuner will tune out discontinuities of 2 to 1 in coaxial transmission line systems or adjust residual VSWR to 1.000 of loads, antennas, etc. May also be used to introduce a mismatch into an otherwise matched system.

M. C. JONES COAXIAL TUNER is designed for extreme ease of operation, with no difficult laboratory techniques involved. Reduces tuning time to a matter of seconds. Graduations on carriage and probe permit resetting whenever reusing the same termination.

SPECIFICATIONS

Impedance	50.0 ohms
Frequency Range	Model 151N 200-1000 Mcs. Model 152N 500-4000 Mcs.
RF Connectors	EIA $\frac{7}{8}$" 50.0 ohm Flange plus adapters to N female connector
Power Rating	100 watts
Range of Correction	VSWR as high as 2 may be reduced to a value of 1.000

FOR MORE INFORMATION ON TUNERS, DIRECTIONAL COUPLERS, R. F. LOADS, Etc., PLEASE WRITE TO:

M.C. Jones

M. C. JONES ELECTRONICS CO., INC.

185 N. MAIN STREET, BRISTOL, CONN.

SUBSIDIARY OF Bendix AVIATION CORPORATION

VOICE *via* LIGHT BEAM

In Space . . .
On Earth . . .

New power sources and electronic controls now permit transmission of voice communications between line-of-sight points using a narrow, high intensity beam of light . . .

FOR

- **Military applications requiring maximum security**
- **Civilian applications requiring privacy and where wires are not available**
- **Use in hazardous locations where radio waves could cause explosions or interference**

Electron Arc's power supplies are designed for use with high intensity light source lamps such as the Sylvania concentrated arc lamps. Complete units are available from 2 to 300 watts with varying light source diameters from 0.005 to 0.110 inch—with average brightness of 15,000 to 30,000 candles per square inch.

When used with the appropriate concentrated arc lamps, the system provides a ready and highly efficient source of *infrared* energy. Such units are finding extensive applications in communication and detection systems.

EA's line of power supplies and accessories for high intensity light sources are ideal for use in:

- **Laboratory instruments**
- **Photographic equipment**
- **Projection systems**
- **Photomicroscopy**
- **Collimating systems**
- **Medical and surgical equipment**

Other EA products include transformers and power equipment for special applications with a wide range of frequencies, and voltages to 100 kw and higher. DC power sources with controlled outputs and special voltages include rectifiers for higher voltages and currents than normally available.

IONICS

ELECTRON ARC DIV.
IONICS, INCORPORATED
244 Broad Street • Lynn, Massachusetts

New LAMBDA Regulated Power Supplies

5 and 10 AMP 0-32 VDC

CONVECTION COOLED

5-Year Guarantee

GUARANTEED FOR 5 YEARS

- **Convection cooled—no internal blowers to wear out**
- **Ambient temperature 50°C**
- **Excess ambient thermal protection**
- **Fast transient response**
- **Special, high purity foil, hermetically sealed long-life electrolytic capacitors**
- **Hermetically sealed transformer designed to MIL-T-27A**
- **Remote sensing and DC vernier**

New LAMBDA LA Series Condensed Data

DC OUTPUT:

(Regulated for line and load)

MODEL	VOLTAGE RANGE[1]	CURRENT RANGE[2]	PRICE
LA50-03	0-32 VDC	0- 5A	$395
LA50-03M	0-32 VDC	0- 5A	$425
LA100-03	0-32 VDC	0-10A	$510
LA100-03M	0-32 VDC	0-10A	$540

[1] The output voltage for each model is completely covered in four steps by selector switches plus vernier control and is obtained by summation of voltage steps and continuously variable DC vernier as follows:

MODEL	VOLTAGE STEPS
LA 50-03, LA 50-03M	2, 4, 8, 16 and ±2 volt vernier
LA100-03, LA100-03M	2, 4, 8, 16 and ±2 volt vernier

[2] Current rating applies over entire output voltage range

Regulation: Line: Better than 0.15 per cent or 20 millivolts (whichever is greater). For input variations from 100-130 VAC. Load: Better than 0.15 per cent or 20 millivolts (whichever is greater).

Transient Response: Line or Load: Output voltage is constant within regulation specifications for step function line voltage change from 100-130 VAC or 130-100 VAC or for step-function load change from 0 to full load or full load to 0 within 100 microseconds after application.

Ripple and Noise: Less than 1 millivolt rms with either terminal grounded.

AC INPUT:

100-130 VAC, 60 ± 0.3 cycles[3]

[3] Well within standard commercial power line frequency tolerances in the United States and Canada.

OVERLOAD PROTECTION:

Electrical: Magnetic circuit breaker front panel mounted. Special transistor circuitry provides independent protection against transistor complement overload. Fuses provide internal failure protection. Unit cannot be injured by short circuit or overload.

REMOTE SENSING:

Provision is made for remote sensing to minimize effect of power output leads on DC regulation, output impedance and transient response.

PHYSICAL DATA:

Size:
LA 50-03 . . . . 5¼" H x 19" W x 14⅜" D
LA100-03 . . . . 7" H x 19" W x 14⅜" D

Panel Finish: Black ripple enamel (standard). Special finishes available to customers specifications at moderate surcharge. Quotation upon request.

Send today for complete data

λ LAMBDA ELECTRONICS CORP.
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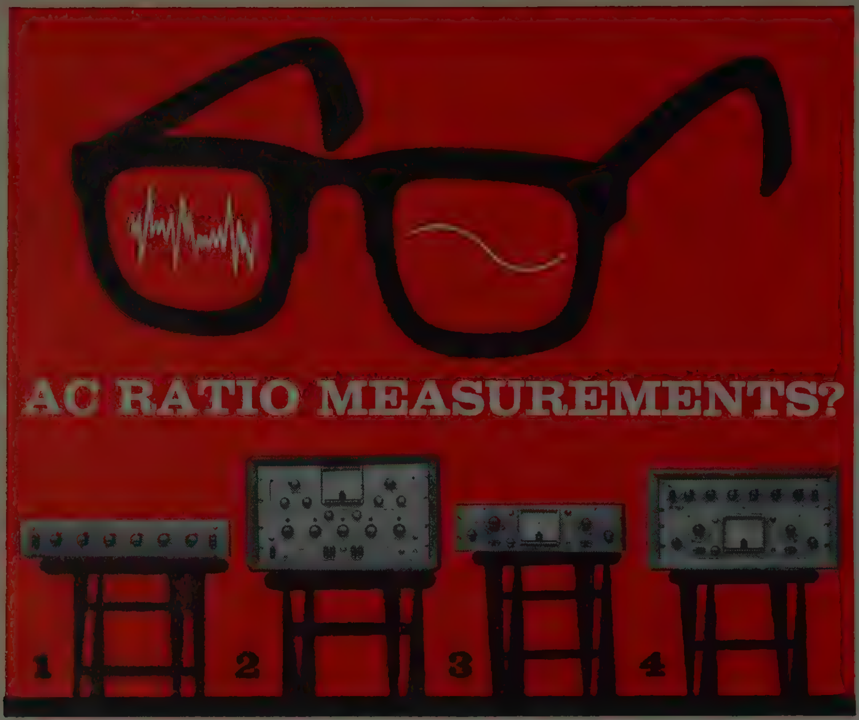


THERE'S A NORTH ATLANTIC INSTRUMENT TO MEET YOUR REQUIREMENTS, TOO...

Now—from North Atlantic—you get the complete answer to AC ratio instrumentation problems—in the laboratory, on the production line, in the field.

Specialists in ratiometry, North Atlantic offers the only complete line of precision instruments to handle any ratio measurement task. All are designed to meet the most demanding requirements of missile age electronics—provide high accuracy, flexibility, component compatability and service-proven perforformance. Some are shown above.

If your project demands total solution to ratio measurement problems, write for Date File No. 10M. It provides complete specifications and application data and shows how North Atlantic's unparalleled experience in ratiometry can help you.

1. RATIO BOXES:	2. COMPLEX VOLTAGE RATIOMETERS	3. PHASE ANGLE VOLTMETERS	4. RATIO TEST SETS
Both laboratory standards and general duty models. Ratio accuracies to 0.0001%. Operation from 25 cps to 10 kc.	Integrated, single-unit system for applications where phase relations are critical. Accuracy to 0.0001%, unaffected by quadrature. Three frequency operation. Direct reading of phase shift in milliradians or degrees.	Versatile readout system for all ratiometry applications, providing direct reading of phase, null, quadrature, in-phase and total voltage. Broadband, single-, or multiple-frequency operation.	Ratio reference and readout in one convenient package for production line and similar applications. Can be supplied with any desired combination of ratio box and phase angle voltmeter.

NORTH ATLANTIC INDUSTRIES, INC.
603 MAIN STREET, WESTBURY, N.Y. • EDGEWOOD 4-1122

TUBE PROBLEM:

The Armed Forces needed a new version of the 6J4 reliable tube type which would provide a tube life of almost 1000 hours. Existing tubes of this type had an average life of only 250 hours. In addition, this new tube had to be produced under ultra-high quality control standards.

SONOTONE SOLVES IT:

By making improvements in the cathode alloy and setting up extremely tight controls in precision, manufacture and checking, Sonotone engineers produced a 6J4WA with a *minimum* life of 1000 hours...most running *much longer.*

RESULTS:

The Sonotone 6J4WA is one of three reliable tubes now being manufactured under U.S. Army Signal Corps RIQAP (Reduced Inspection Quality Assurance Program), monitored by the U.S. Army Signal Supply Agency. And the same rigid quality standards apply to Sonotone's entertainment type tubes as well.

Let Sonotone help solve *your* tube problems, too.

Sonotone® CORP.
Electronic Applications Division, Dept. T42-40
ELMSFORD, NEW YORK
Leading makers of fine ceramic cartridges, speakers, microphones, tape heads, electron tubes.
In Canada, contact Atlas Radio Corp., Ltd., Toronto

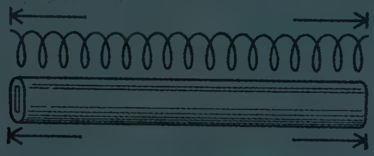
NUMBER 4
RELIABILITY SERIES

BOURNS TRIMPOT® WITH BUILT-IN TEMPERATURE STABILITY

Stable settings under extreme temperature conditions is an outstanding feature of the Trimpot® potentiometer. This thermal stability is built-in through all phases of design and production—

MATCHED COEFFICIENTS OF THERMAL EXPANSION

Resistance wire and mandrels have matched coefficients of thermal expansion to reduce the "strain gage effect." Linear expansion rates for the mandrel and wire match so closely that the temperature coefficient value for the entire wirewound element approximates that of the wire itself.

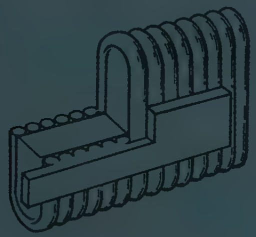


THERMALLY STABLE CERAMIC MANDRELS

Bourns takes advantage of high thermal stability of ceramic materials for element mandrels. Today, all Bourns Trimpot potentiometers provide the improved performance and reliability afforded by ceramic materials.

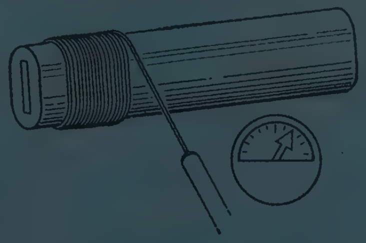
EXCLUSIVE SILVERWELD® TERMINATION

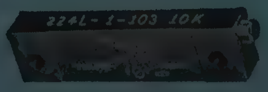
Silverweld is an actual metal-to-metal fusion of element wire and external terminal. In doing away with mechanical or soft-solder joints, Bourns eliminates potential hot spots thus extending the potentiometer's temperature range. The fusion of the Silverweld terminal to many turns of wire on the resistance element avoids the problem of single wire termination. Silverweld is virtually indestructible under thermal stresses.



EXCLUSIVE TENSION CONTROL EQUIPMENT

Bourns has developed specialized winding equipment that provides constant and precise control of wire tension during winding operations. "Necking" of the wire or resistance-altering stresses never occur. Instead the wire remains uniform—well able to withstand temperature variations with no appreciable change in resistance.





Specify Trimpot — the original leadscrew-actuated potentiometer with reliability on which you can depend. 20 basic models — 4 terminal types — 3 mounting styles.

BOURNS

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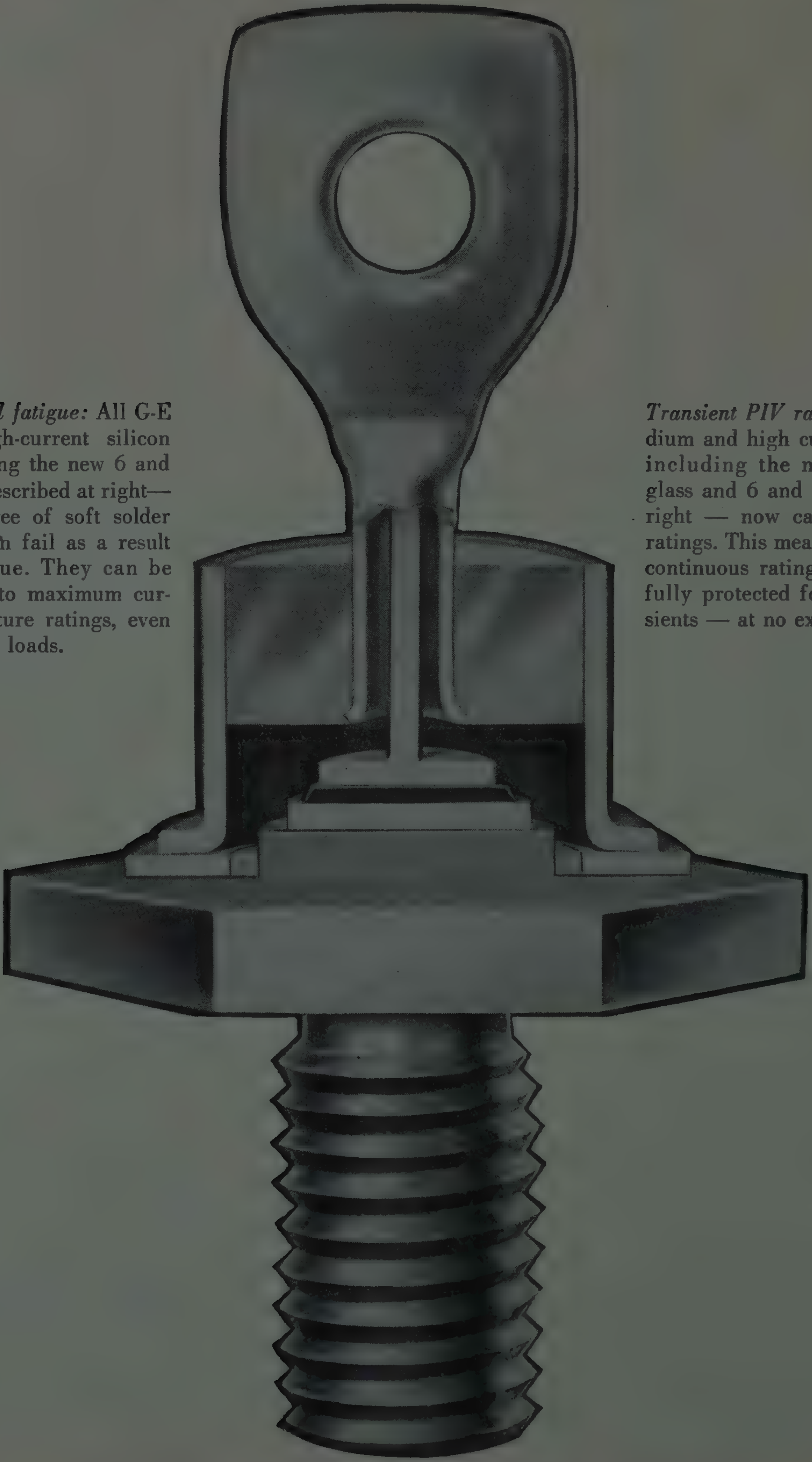
PLANTS: RIVERSIDE, CALIF. AND AMES, IOWA

Write for new Trimpot summary brochure and list of stocking distributors.

Exclusive manufacturers of Trimpot®, Trimit® and E-Z-Trim®. Pioneers in transducers for position, pressure and acceleration.

NEW RECTIFIERS FROM

Free from thermal fatigue: All G-E medium and high-current silicon rectifiers—including the new 6 and 12-amp. devices described at right—are completely free of soft solder joints, which often fail as a result of thermal fatigue. They can be worked right up to maximum current and temperature ratings, even on highly cyclical loads.



Transient PIV ratings. All G-E medium and high current rectifiers — including the new subminiature glass and 6 and 12-amp. devices at right — now carry *transient* PIV ratings. This means you can buy the continuous rating you need and be fully protected for occasional transients — at no extra cost!

GENERAL ELECTRIC

New Silicon Subminiature Glass Rectifier

Designed for maximum thermal conductance over a wide temperature range. Suitable for MIL-E-1/1143. Extremely low leakage currents. Ideal for magnetic amplifier, blocking and other low-leakage applications.

JEDEC or GE Type Number	Repetitive PIV	Transient PIV	Max. I_{DC} at T°C	Max. Lkge Cur. (Full cycle Av.)	Max. Full Load Voltage Drop (Full cycle Av.)	Max. Oper. Temp.
			@ 150°C	@ 100°C	@ 25°C	
1N645	225	275	150 ma	15 μa	1V	175°
1N646	300	360	150 ma	15 μa	1V	175°
1N647	400	480	150 ma	20 μa	1V	175°
1N648	500	600	150 ma	20 μa	1V	175°
1N649	600	720	150 ma	25 μa	1V	175°
				@ 150°C		
1N677	100	20	400 ma	.2 ma	1V	175°

1N676-1N679, 1N681-1N687 and 1N689 also available in this package.

New Silicon Insulated Stud Mounted Junction Rectifier

Designed for applications requiring fins or direct chassis mounting. Stud electrically insulated from rectifying junction. High forward currents permitted at case temperatures up to 150°C (up to 165°C with derating). Reverse current at maximum junction temperature extremely low, making these devices ideal for low-leakage applications.

JEDEC or GE Type Number	PIV	Max. I_{DC} at T°C	Max. Peak 1 cycle Surge	Max. Lkge Cur. (Full Cycle Av.)	Max. Full Load Voltage Drop (Full Cycle Av.)	Max. Oper. Temp.
		@ 50°C Case		@ 150°C	@ 150°C	
1N2851	500	1.5 Amps	15 Amps	.3 ma	.65V	150°
1N2852	600	1.5 Amps	15 Amps	.3 ma	.65V	150°
		@ 75°C Case				
1N2847	100	1.5 Amps	15 Amps	.4 ma	.65V	165°
1N2848	200	1.5 Amps	15 Amps	.3 ma	.65V	165°
1N2849	300	1.5 Amps	15 Amps	.3 ma	.65V	165°
1N2850	400	1.5 Amps	15 Amps	.3 ma	.65V	165°

New Silicon Medium Current 6 and 12-amp. Junction Rectifiers

With these new devices, General Electric now offers the widest selection of rectifiers in the medium current range. Designed for all rectifier applications from 2 to 15 amperes. Extremely low forward voltage drop and thermal impedance combined with high junction temperature rating permit high current operation with minimum space requirements. May be mounted directly to chassis or fin or electrically insulated from heat sink by using mica washer kit provided with each unit.

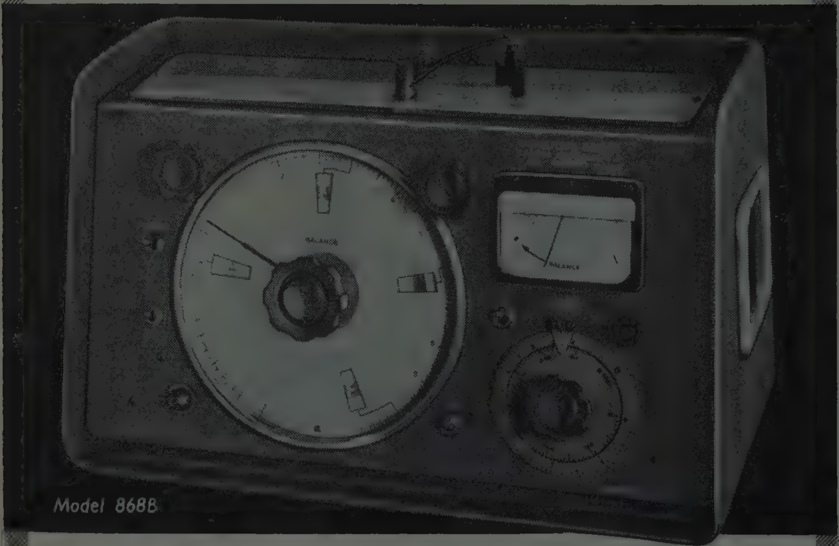
JEDEC or GE Type Number	Repetitive PIV	Transient PIV	Max. I_{DC} @ 145°C Stud Single Phase	Max. Peak 1 Cycle Surge	Max. Lkge Cur. (Full cycle Av. @ Full Load)	Max. Full Load Voltage Drop (Full Cycle Av.)	Max. Oper. Temp. °C
					@ 150°C Stud	@ 150°C Stud	
1N1341A	50	100	6A	150A	3 ma	.64V	200°
1N1342A	100	200	6A	150A	2.5 ma	.64V	200°
1N1343A	150	300	6A	150A	2.25 ma	.64V	200°
1N1344A	200	350	6A	150A	2.0 ma	.64V	200°
1N1345A	300	450	6A	150A	1.75 ma	.64V	200°
1N1346A	400	600	6A	150A	1.5 ma	.64V	200°
1N1347A	500	700	6A	150A	1.25 ma	.64V	200°
1N1348A	600	800	6A	150A	1.0 ma	.64V	200°
1N1199A	50	100	12A	240A	3 ma	.55V	200°
1N1200A	100	200	12A	240A	2.5 ma	.55V	200°
1N1201A	150	300	12A	240A	2.25 ma	.55V	200°
1N1202A	200	350	12A	240A	2.0 ma	.55V	200°
1N1203A	300	450	12A	240A	1.75 ma	.55V	200°
1N1204A	400	600	12A	240A	1.5 ma	.55V	200°
1N1205A	500	700	12A	240A	1.25 ma	.55V	200°
1N1206A	600	800	12A	240A	1.0 ma	.55V	200°

For more information, see your General Electric Sales Representative, or write Semiconductor Products Dept., Electronics Park, Syracuse, N. Y. In Canada: Canadian General Electric Co., 189 Dufferin St., Toronto, Ont. Export: International General Electric Co., 150 E. 42 St., N.Y.C. See your authorized General Electric Distributor for fast service, factory-low prices.

GENERAL ELECTRIC

THE NEW MARCONI UNIVERSAL BRIDGE

Gives NEW Simplicity in LCR Measurements



Model 868B

L — 1 μH to 100 henrys
C — 1 μμF to 100 μF
R — 0·1 ohm to 100 MΩ

•Direct read-out with no multiplying factors, eliminates operator errors. •Model 868 B also has precision Q and tan δ (D) dials. Inductance and capacitance are measured at 1 or 10 kc/s in an R-C ratio-arm bridge; resistance at d.c. in a Wheatstone bridge. The bridge detector gives positive indication of the direction of balance point even when far off-balance; as a result, components whose values are completely unknown can be evaluated in a few seconds with the minimum of searching. Detector a.g.c. eliminates the need for sensitivity controls.
Also available—**Low Capacitance Bridge Model 1342:** 0·002μμ F to 1,111μμ F; 3-terminal transformer ratio-arm bridge designed for precision measurement of extremely low capacitance.
For full details, write for leaflet D171.

MARCONI INSTRUMENTS

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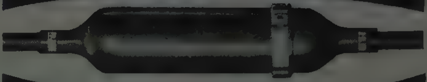
ROBERTSON

SPLICE CASES—made of polyvinyl chloride, for power or communication cable.



- designed for fast, convenient splicing, saving hours of time and labor.
- provide permanent moisture and dust-proof seal.

CONNECTOR CASES—maximum protection for connectors used in field equipment.
- fit any size or type of connector.
- stepped ends can be cut to size of wire or cable being used.

For further information write:
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. . . permits instantaneous line transient absorption, protects transistor circuitry and provides superior performance. Models up to 100 volts and 10 amperes. Write for literature R3

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Horizons Unlimited

Development of a high-performance inertial guidance system of unprecedented performance for long-range guided missiles, satellites and space vehicles has been announced by Bell Aircraft Corporation's Avionics Division.

Bell Avionic's engineers describe the highly-classified system as "the most successful and reliable of any new inertial instrumentation concepts so far tested."

The system was developed under the direction of Dr. Helmut W. Schlitt, recognized within the industry as an outstanding authority in the field of inertial guidance.

The new system has undergone extensive flight tests at the Niagara Falls, N.Y., Municipal Airport. Some of its components already are being used in guided missiles.

Bell Aircraft's Avionics Division also is engaged in many other important programs, including Battlefield Monitoring Systems, All-Weather Automatic Aircraft Landing Systems and Secure Data Link Systems.

The Avionics Division recently has joined with Pan-American World Airways, Inc. to find a solution to the serious problem of radio interference with the operation of Army combat electronic equipment.

Bell and Pan-Am's Guided Missiles Range Division will set up an electronic environmental test facility and drone test range near Fort Huachuca, Ariz., to track down the source of the interference, submit recommendations for corrective action and initiate procedures to eliminate these conditions.

Experienced engineers and scientists are needed for all of these programs. For information, write George Klock, director of engineering employment, today.

AVIONICS DIVISION
BELL
AIRCRAFT
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BUFFALO 5, NEW YORK

RADIO RESEARCH INSTRUMENT CO.

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F-28/APN-19 FILTER CAVITY

Jan. spec: Tuneable 2700-2900mc, 1.5db max. loss at ctr freq over band. Details: Insertion loss variable. Single tuned filter for freq channelling in radar beacon. Invar center tuning conductor ¾ wavelength. New $37.50 each.

2C40 LIGHTHOUSE CAVITY

AN/APW-11A transmitter cavity for 2C40. Complete S band coverage at max. power. Temperature compensated. New $77.50 ea.

AN/APS-10 3CM. X BAND RADAR

Complete RF head including transmitter, receiver, modulator. Uses 2J42 magnetron. Fully described in MIT Rad. Lab. Series Vol. I, pps. 616-625 and Vol. II, pps. 171-185. $375.00. Complete X band radar system also avail. incl. 360 deg. antenna, PPI, sync, pwr supply. Similar to $17,000 weather radar now in use by airlines $750 complete.

10 CM. WEATHER RADAR SYSTEM

US Navy Raytheon 275 KW peak output S band. Rotating yoke Plan position Indicator. Magnetron supplied for any S band frequency specified, incl. Weather Band. 4, 20 and 80 mile range. 360 degree azimuth scan. Sensitive rcvr using 2K28/707B and 1N21B. Supplied brand new complete with instruction books and installation drawings. Can be supplied to operate from 32VDC or 115 volts. Price $975. Ideal for weather work. Has picked up clouds at 50 miles. Weight 488 lbs.

SUMMATION BRIDGE

TS 731/URM. Portable, general purpose, field type, ruggedized wattmeter used to measure R.F. power output of radio, radar & navigation equips. A direct reading summation type circuit in which a bolometer acts as one arm of a D.C. Weatstone bridge, at the same time providing correct resistivity termination for R.F. entering the wattmeter. Power supply: 115-VAC, 1 ph.50-1600 cy. Freq. range: 1000 to 4000 mcs with plumbing supplied. VSWR less than 1.3 Gov't cost well over $1000 in large quantity. Your cost—$395 ea. Factory new.

AN/APN-60

10 CM. RADAR BEACON FOR GUIDED MISSILES. 14″ pressurized housing. 2C40 lighthouse OSC. Trans-revr. unit $275.00.

⅞″ RIGID COAX. RG44/U 50 ohm, standard fittings. 10cm stub supported. 12 ft. lengths. Silver plated. New $34.50 each. 12 ft. length. Right angle bends $6 ea.

X BAND HORN. Waveguide feed pressurized window. 180 Deg. feed back. New $22.50.

TOPWALL HYBRID JUNCTION. 8500-9600mc 1x.5 wg size. Broad banded better than 16%. Aluminum casting. $15.00 new. Crossover output. 1x.5 wg size. $5.00 new.
BROAD BAND BAL MIXER using short slot-hybrid. Pound type broad band dual balanced crystal holder. 1x.5 wg size. $25.00 new.

FLEXIBLE WAVEGUIDE. 1x.5 X band 4″, new $5.00. 1x.5 X band 9″ Technicraft. New $10.00. 1.5 X band 24″ Airtron. New $21.50. 1¼″ x ⅝″ X band 12″ Western Elec. New $19.50.
RG48 TO ⅞ COAX. ADAPTER S Band 1½ x 3″ W.G. to RG 44/U coax. New. $26.50.
COAX MIXER ASSEMBLY S BAND IN21 type crystal detector RF to IF, 'N' fittings, matching Slug, duplex couplings, mf. G.E. New, $18.50.

3 & 10 CM COMPLETE SCR 584 RADARS

AUTOMATIC TRACKING RADAR

Our 584s in like new condition, ready to go, and in stock for immediate delivery. Ideal for research and development, airway control, GCA, missile tracking, balloon tracking, weather forecasting, antiaircraft defense, tactical air support. Write us. Fully Desc. MIT Rad. Lab. Series, Vol. I, pps. 207-210, 228, 284-286.

AN/MPN-1B GCA SET

Ground control approach equipped trailer with 3 cm precision and 10 cm search radars plus full complement precision and search indicators. All in original trailer. Call—write for info. and price. Fully Desc. MIT Rad. Lab. Series, Vol. II, pps. 237-251.

AN/FPN-32 GCA RADAR

Lab. for Electronics "Quad" type portable ground control approach radar system. 3cm. search and precision approach. Complete systems in used, good condition. A very late type system—in stock $5500 ea.

MIT Model 9 PULSER
1 MEGAWATT-HARD TUBE

Output pulse power 25KV at 40 amp. Max duty ratio: .002. Uses 6C21 pulse tube. Pulse duration .25 to 2 microsec. Input 115 volts 60 cycle AC. Includes power supply in separate cabinet and driver. Fully guaranteed as new condition. Full Desc. MIT. Rad. Lab. series "Pulse Generators."

24KMC. PKG. MAGNETRON

3J21 Magnetron. 60kw output at 1.25cm, K band. 15kv 15 amp. input. Axial cathode mount. Waveguide output. Packaged with magnet. New $87.50 ea. gtd.

SPERRY KLYSTRONS

SMX-32 Amplifier-Multiplier 9.0-10.5-KMC. Amp. freq. multiplier; output power 1.5 to 2 watts at 9000 to 10,500 mc; drive freq. 4500 to 5250 mc. New in original sealed cartons. Guaranteed 90 days. Regular price $925 ea. Our price $425 ea.

SMC-11A Frequency multiplier. Output range 4640-4670 mc. ½ to 1 watt on 6th harmonic. Input range 773-778 mc. For direct control of microwave frequencies. New in original sealed cartons. Guaranteed 90 days. Regular price $1500 ea. Our price $495 ea.

2 MEGAWATT PULSER

Includes Rectifier xfmr 5800/7000V, 2.2/2.55-KVA; Resonant charging choke 150 cy. 30H 12 Amp. Ins. 17KV; Capacitor network 17E2-2-300-25P2T; Capacitor .04Mfd. 17KVDC; Transformer 4400V to 22,000V; Fil. Xfmr. & Filter choke 5.1V 18Amp, 6 H .21Amp. DC; 4C35 pulser, 3B24W (6 each), WL1B41 gap tube. Price new type MD-53/APS20C $1125.00

AN/SPT-6A TRANSMITTER

350 to 1400mc countermeasures jammer. 115V 60 cy AG input. New and complete operating unit. $750.00

FM 3CM. SIGNAL GENERATOR

TS-263/TPS-10 X band FM signal generator. Tunes 9KMC band. With calib. atten. Regulated pwr supply. New $325.00
TAPER. RG51 to RG52 (1¼ x ⅝″ to 1 x ½″) Smooth Electroform. Standard Flankes. New $16.50.

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Complete description in McGraw-Hill Radiation Laboratory Series, Volume I, page 284 and page 209, and Volume 26, page 233.

RT39/APG-5 & 15 10CM RADAR. Complete S band RF package. Lighthouse 2C40 xmtr. 2C43 rcvr, TR. 829B pulser, miniature 6AK5 IF strip. Press. 12″ dia., 24″ lg. New with tubes $275. Ref: MIT Rad. Lab. Series Vol. I, pg. 207.

74 to 320 MC BUTTERFLY TUNER

Rotates thru 360 deg. either hand or motor driveable. New, $22.50 ea.

VG-12″ PROJECTION PPI RRTR $395.00.
CPS-1, PPI 12″. NEW. $275.00.

TS-743/U SIGNAL GENERATOR

15,250 to 16,250mcs, calibrated attenuator, 115VAC regulated power supply. Mfg. Polarad. Exc. condition $1250. each.

KLYSTRON MOUNTS

3CM. Precision Tube Mount. Waveline model 688. X band shielded klystron mount PRD signal generator type. Complete with variable glass vane attenuator. Brand new. $205. list. Price $45.00.
S. band. Type N output. Tunable over entire band. For Shepard type tube i.e. 726 w/socket & tube clamp. Mfg. GE. New, $15.00.
K band. 2K50 output cplg. w/90 deg. H bend. New. $19.50.

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

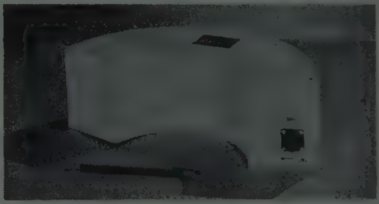
(Continued from page 183A)

any machine requiring two- or three-axis positioning, such as drilling, boring, turning equipment, punch presses, spot welders, and eyelet inserters for circuit board sub-assemblies.

Copies of the brochure may be obtained by writing to the firm.

"Shaped" Battery With Dual Voltage

A special, "shaped" electric APU power source containing two separate battery sections developed by **Cook Co.,** a subsidiary of **Telecomputing Corp.,** 3850 Olive St., Denver 7, Colorado, provides dual output for missile and spacecraft power requirements. The two battery sections in the Model P68A provide two different voltage levels.



One section provides a current of 8 amperes at 28 volts. Maximum current is 25 amperes, with a discharge time of 40 minutes at 8 amperes. Capacity is 5.5 ampere-hours.

The second section supplies 6.3-volt power at 3 amperes. Discharge time is 40 minutes. Maximum current is 25 amperes. Capacity is 5.5 ampere-hours.

Both sections are activated automatically. The activating mechanism is an electrolyte tank and piston arrangement, operated by a solid propellant gas generator. Activation time is 1 second. The signal required is 2 amperes at 28 volts.

The cells are specially designed for automatic activation and have foil anodes enclosed in sealed separators. This technique eliminates the problems of anode gassing and internal shorts. The entire unit is hermetically sealed to prevent leakage or contamination. The battery weighs 13 pounds.

The Model P68A will withstand shock to 50 g, acceleration to 20 g and vibration to 10 g, along all three major axes. Temperature range is from 50°F to 150°F, with special models providing ranges of $-65°F$ to $+165°F$. Thermostatically controlled heaters are available.

Case material is steel. Mounting configuration is supplied to customer specifications. Standard or special connectors and terminals may be specified.

A fail-safe detection circuit may be provided for protection against premature activation. A double estimated activation ignition circuit overcomes shorts or open circuit failures. The estimated shelf life is 5 years.

(Continued on page 196A)

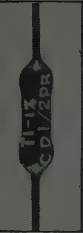
How to determine high-frequency characteristics of precision film resistors

Specify with confidence from this complete line of time-proved TI resistors

MOLDED†

TI type number	wattage rating watts	MIL designation	standard resistance ranges	max. recommended voltage volts
CDM⅛	⅛	RN60B	10 Ohm-1 Meg	350
CDM¼	¼	RN65B	10 Ohm-1 Meg	500
CDM½	½	RN70B	10 Ohm-5 Meg	750
CDM 1	1	RN75B	10 Ohm-10 Meg	1000
CDM 2	2	RN80B	50 Ohm-50 Meg	2000

MIL-LINE †



TI type number	wattage rating watts	MIL designation	standard resistance ranges	max. recommended voltage volts
CD⅛R	⅛	—	10 Ohm-1 Meg	350
CD¼R	¼	RN10X	10 Ohm-1 Meg	500
CD½PR	½	RN15X	10 Ohm-3 Meg	650
CD½MR	½	RN20X	10 Ohm-5 Meg	750
CD½SR	½	—	50 Ohm-10 Meg	850
CD1R	1	RN25X	10 Ohm-10 Meg	1000
CD2R	2	RN30X	50 Ohm-50 Meg	2000

HERMETICALLY SEALED LINE †

TI type number	wattage rating watts	MIL designation	standard resistance ranges	max. recommended voltage volts
CDH⅛M	⅛	—	10 Ohm-500K	250
CDH⅛	⅛	RN60B	10 Ohm-1 Meg	350
CDH¼	¼	RN65B	10 Ohm-1 Meg	500
CDH½P	½	—	10 Ohm-3 Meg	650
CDH½A	½	RN65B	10 Ohm-3 Meg	650
CDH½M	½	RN70B	10 Ohm-5 Meg	750
CDH½S	½	—	50 Ohm-10 Meg	850
CDH 1	1	RN75B	10 Ohm-10 Meg	1000
CDH 2	2	RN80B	50 Ohm-50 Meg	2000

†All values available in 1% tolerance; nominal lead length 1.5 in.

sensistor* SILICON RESISTORS

Standard available resistances @ 25° C:
68, 82, 100, 120, 150, 180, 220, 270, 330, 390, 470, 500, 560, 680, 820, 1000, 1200, 1500, 1800 ohms.‡

Type No.	Wattage Rating	Body Dimensions		Average Temperature Coefficient	Resistance Tolerance
	W	Length	Diameter	%/°C	%
TM ⅛	⅛	0.585″	0.200″	+0.7	±10
TM ⅛	⅛	0.406″	0.140″	+0.7	±10
TC⅛	⅛	TO—5	Transistor	+0.7	±10

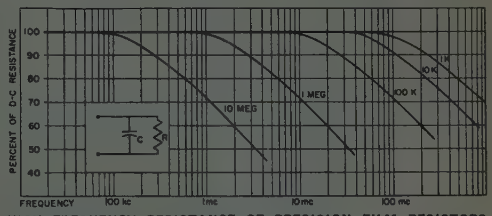
* TRADEMARK OF TEXAS INSTRUMENTS INCORPORATED
‡ Other resistance values and tolerances available on special order

For a more detailed discussion of this subject, contact your nearest TI sales office for a copy of "High-Frequency Characteristics of Precision Film Resistors."

In high frequency applications, precision film resistors are superior to composition or wirewound resistors; skin effect of the thin film is negligible.

OHMIC VALUE vs FREQUENCY

Precision film resistors of a given physical size have the same distributed capacitances regardless of their ohmic value. As the frequency increases, the shunting effect of the distributed capacitance causes the effective parallel resistance to decrease. The reactance of the stray capacitance becomes a relatively good shunt when it approximates the ohmic value of the resistor. The smaller the ohmic value of a precision film resistor (for a given physical size), the higher its usable frequency range.



HIGH FREQUENCY RESISTANCE OF PRECISION FILM RESISTORS

INDUCTANCE CONSIDERATIONS

The inductance caused by helixing the higher value resistors is negligible throughout the "useful" range of frequencies at which the resistance is greater than 60% of its d-c value.

When resistors under 500 ohms are measured using high frequency meters, the reactive component of the equivalent parallel circuit appears inductive because of lead and binding post inductance. However, the resistor itself is capacitive.

TI TYPE	SIZE (WATT RATING)				
	⅛	¼	½	1	2
MIL-LINE (CD)	0.2	0.1	0.25	0.5	0.6
MOLDED (CDM)	0.3	0.25	0.45	0.7	0.7
HERMETICALLY SEALED (CDH)	0.3	0.25	0.45	0.75	0.8

CAPACITANCE IN $\mu\mu$f OF TI PRECISION FILM RESISTORS

CAPACITANCE CONSIDERATIONS

The average measured capacitance of Texas Instruments Precision Film Resistors is determined primarily by the end cap-to-cap capacitance which is proportional to the dielectric constant of the core and encapsulating material.

MOUNTING

Precision film resistors of 200 ohms or less perform satisfactorily at 5000 mc and higher if placed in a well-designed coaxial mount. A coaxial mount constructed from a standard UG-18B/U Type N plug can be used effectively. In conventional terminals, correct mounting of the body of the resistor off the circuit chassis and the use of short leads will minimize the stray capacitance and lead inductance.

Specify TI precision resistors!

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NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 194A)

Correction Notice

Delay Lines

By using a new principle in delay network design, a new series of delay lines with ratio of rise time to total delay less than 0.02 has been achieved by **Ad-Yu Electronics Lab., Inc.,** 249–259 Terhune Ave., Passaic, N. J. Besides the feature of producing very fast rise time, the attenuation can be minimized to be less than 0.02 db per 1/10-microsecond delay. Two or more units of Type 10T series can be connected in tandem for longer delay whenever their impedances are identical. In this case, the time delay will be equal to the sum of the delay of each unit and the rise time will be approximately equal to the square root of the sum of the squares of the rise time of each unit.

The features of this delay line series can be summarized as follows: (1) The ratio of rise time total delay can be made less than 0.02. (2) The distortion is generally less than 2%. (3) The attenuation can be made less than 0.2 db per microsecond delay. (4) Physical size can be made less than 2 cubic inches per microsecond delay. (5) The temperature coefficient is less than 50 parts per million per degree Centigrade. (6) All types of this series can also be made to meet existent MIL specifications. The characteristic impedance ranges from 50 ohms to 1000 ohms. The time delay per 10 sections ranges from 0.25 microsecond to 5 microseconds. The accuracy of delay is less than ±1.5%.

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1	$4.20	$5.45	$6.15	$7.90	$30.00	37.90	$2.90	$3 .45	$29.75	$36.00
2	5.30	6.75	7.60	9.90						
3	7.60	15.40	17.10	20.80						
6	9.75	27.00	30.00	40.00						
10	19.90	33.60	38.45	46.75						
24	22.00	27.75	33.45	47.15						
50	32.00	55.75	60.45	88.90						
100	84.15	110.30	130.60	174.75						
150	95.45	136.75	180.30	321.90						
200	149.00	170.00	306.00	350.00						
400	241.00	261.00	495.00	561.00						

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60	$57.45	$84.75	$108.60	$57.00	$28.60
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200	189.45	243.75	299.15	189.00	94.60
400	274.90	479.45	585.75	274.00	137.30
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1000	963.15	1359	1620	963.00	513.90

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1	$1.40	$2.30	$3.45	$4.90	$8.75	$10.65	$16.00	
2	2.30	3.20	5.25	6.25	11.70	14.00	21.90	$2.90
4	3.65	7.35	11.40	16.00	24.85	28.50	47.60	3.60
6	4.40	8.75	12.75	18.75	32.60	36.00	63.45	4.70
12	8.00	15.75	21.95	30.95	45.00	52.50	103.60	7.60
20	12.90	25.30	37.45	49.90	78.75	90.00	169.00	9.45
24	16.00	32.00	44.60	57.50	85.00	96.50	188.75	12.75
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18	2.25	2.90	3.45	3.85
35	2.75	3.25	3.90	4.45

DC AMP	400 PIV	500 PIV	600 PIV	700 PIV
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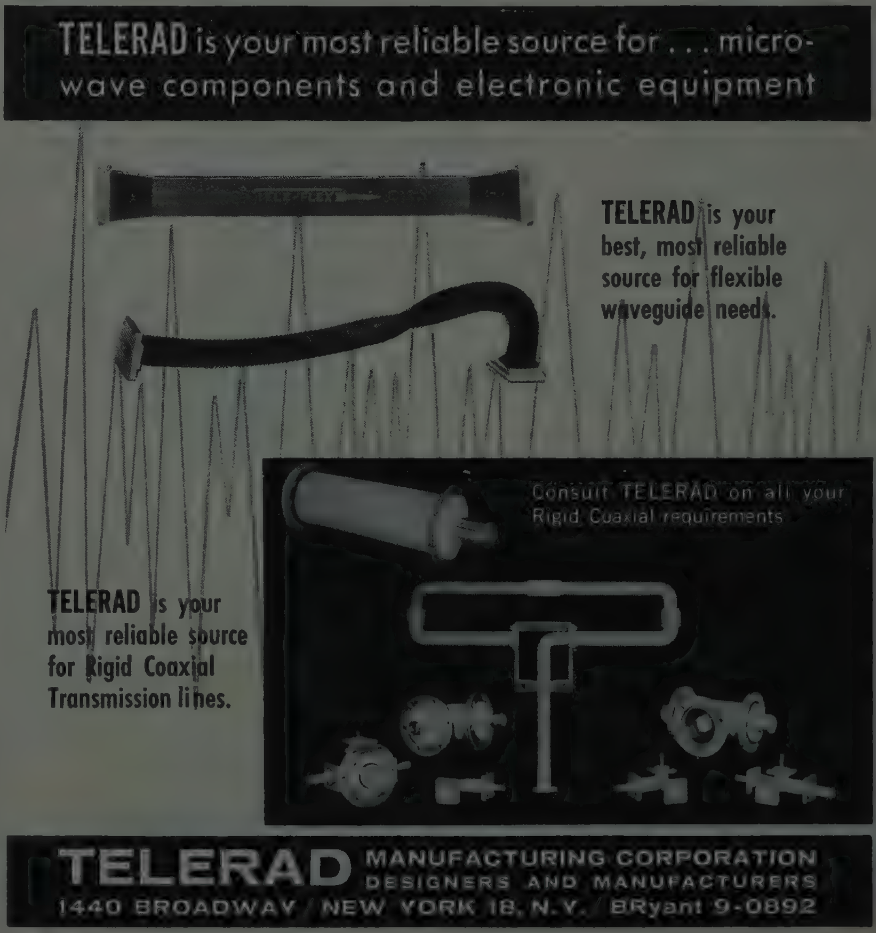

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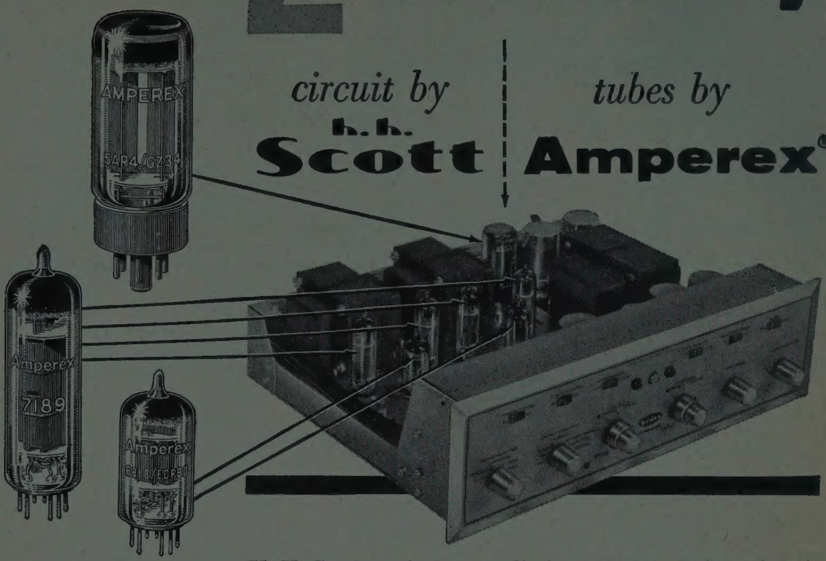

H. H. Scott engineers, preliminary to the design of their *Model 299 (40 Watt) Complete Stereo Amplifier,* canvassed the industry for tube types offering something truly exceptional in the way of reliability, low distortion, low noise, low hum and absence of microphonics.

As has frequently been their experience, the people at Scott found these qualities best exemplified by Amperex tubes. Thus, the tube complement of the Scott Model 299 includes four Amperex 7189's, one Amperex 5AR4/GZ34, and two Amperex 6BL8/ECF80's.

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6BQ5/EL84: 17 w., push-pull
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VOLTAGE AMPLIFIERS
6267/EF86: Pentode for pre-amps
12AT7/ECC81:, **12AU7/ECC82:**, **12AX7/ECC83:** } Twin triodes, low hum, noise and microphonics
6BL8/ECF80: High gain, triode-pentode, low hum, noise and microphonics

RF AMPLIFIERS
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6ER5: Frame grid shielded triode
6EH7/EF183: Frame grid pentode for IF, remote cut-off
6EJ7/EF184: Frame grid pentode for IF, sharp cut-off
6AQ8/ECC85: Dual triode for FM tuners
6DC8/EBF89: Duo-diode pentode

RECTIFIERS
6V4/EZ80: Indirectly heated, 90 mA
6CA4/EZ81: Indirectly heated, 150 mA
5AR4/GZ34: Indirectly heated, 250 mA

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1M3/DM70: Subminiature "exclamation" pattern

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2N1516: RF transistor, 70 mc
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IN542: Matched pair discriminator diodes
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This compact, lightweight grid-dip meter is available in the frequency ranges indicated.

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MOD

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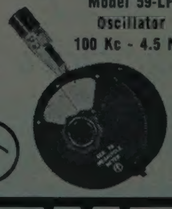
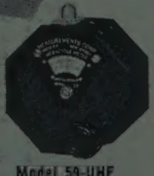
2.2 Mc - 420 Mc


Model 59-LF
Oscillator
100 Kc - 4.5 Mc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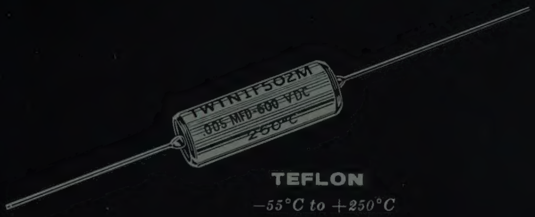

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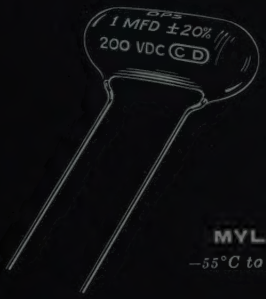

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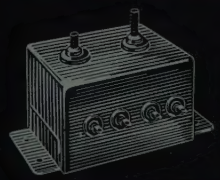

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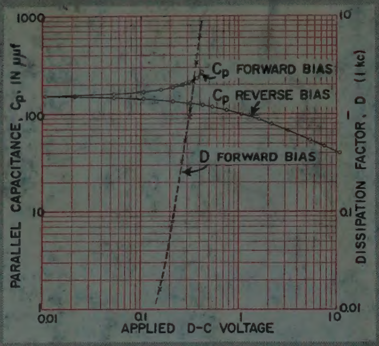
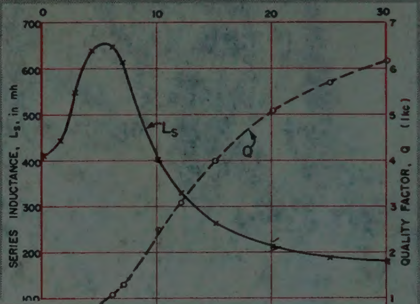
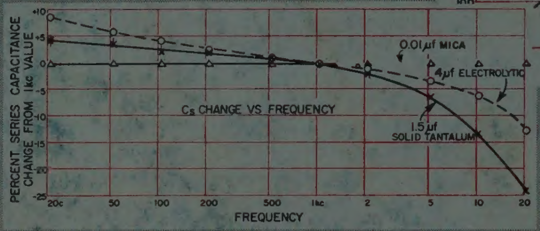


◀ Characteristics of a variable-capacitance diode at 1 kc with forward and reverse biasing.

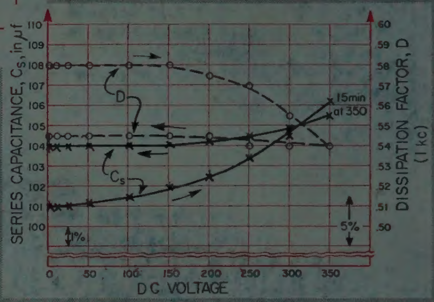


◀ Behavior of an iron-core inductor at various d-c levels. Measurements made with the Bridge in its "Orthonull" position to eliminate sliding balances which occur when making low-Q measurements.

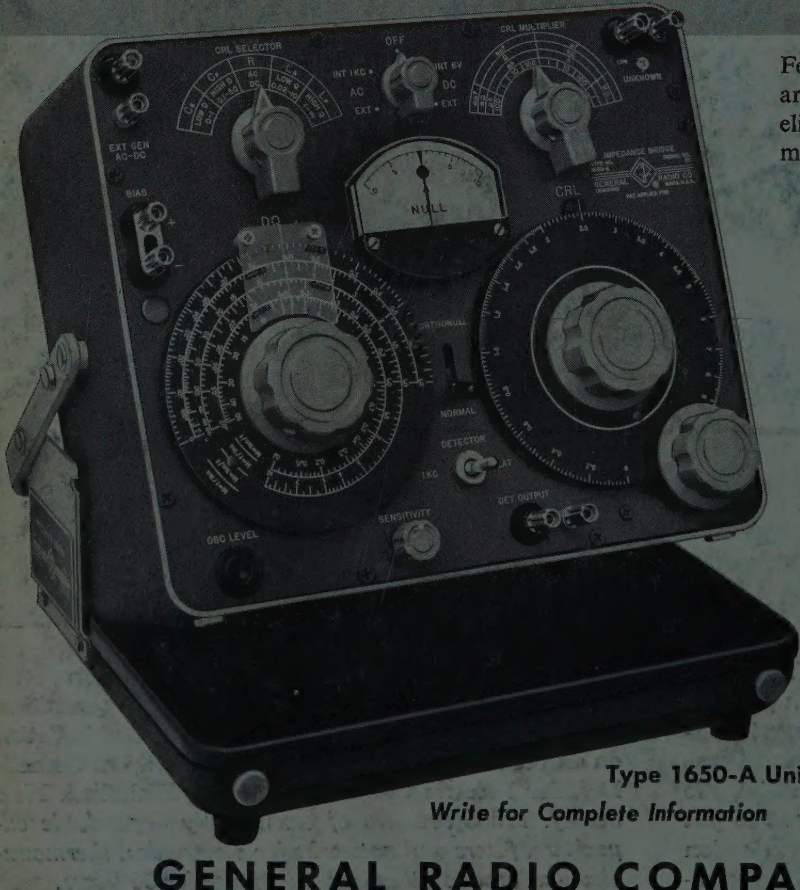


Plot of capacitance change vs frequency for three types of capacitors. External generator used is a G-R Type 1210-C Unit R-C Oscillator. ▲

Variation of capacitance and dissipation factor of an unformed electrolytic as a function of d-c voltage as it is varied to rated voltage and then returned to zero. ▶



These Measurements show why it's called a *Universal* IMPEDANCE BRIDGE



Features ***Orthonull,**** a unique mechanical ganging arrangement of the Bridge's variable elements that eliminates sliding balances when making low-Q measurements.

***U. S. Patent 2,872,639**

- Completely self contained with built-in 1-kc oscillator and selective null detector . . . powered by four "D" batteries . . . total drain is less than 10 ma, providing 1-year battery life for typical laboratory use.
- Useful for measurements from 20c to 20 kc with external generator.
- Unique cabinet design allows panel to be tilted to any convenient angle . . . closes and becomes a rugged carrying case for complete protection.
- Provision for applying polarization voltages to capacitors, biasing voltages to diodes, and small currents to inductors for measurements at various d-c levels.
- **WIDE RANGES** —
  R: 0.001Ω to 10 MΩ **L:** 1 μh to 1000h
  C: 1 μμf to 1000 μf **D:** 0.001 to 50 at 1 kc
  Q: 0.02 to 1000 at 1 kc
- **ACCURACY** —
  1% for L and C from 20c to 20 kc
  1% for R from 20c to 5 kc
  5% for D and Q (ranges are a function of frequency)

Type 1650-A Universal Impedance Bridge . . . $450

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